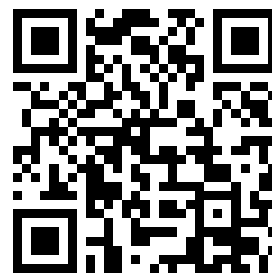


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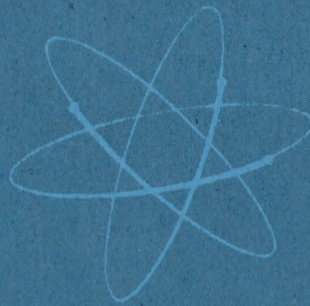
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AF MANUAL 52-8  
VOLUME II

# ELECTRONIC CIRCUIT ANALYSIS



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## Technical Training

### ELECTRONIC CIRCUIT ANALYSIS

*This manual, published in two volumes, of which this is Volume II, provides the graduate electronics repairman or technician with reference information on the fundamentals of electronics. It also makes available a basic coverage of the subject to the student repairman. A prerequisite to understanding these basic concepts is a knowledge of mathematics and electricity. Students who are not too familiar with these subjects may have to consult other sources of information for greater detail. Both Volume I and Volume II include the same attachments of Symbols and Formulas, and Tables, preceding the Index. The Index pertains only to that volume in which it appears.*

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Comments or recommendations for the improvement of this manual should be sent to Headquarters Air Training Command, Randolph Air Force Base, Texas.



# Modulation, Detection, and Single Sideband Transmission

In radio transmission, the steady r-f voltage wave generated by a transmitter alone cannot produce an intelligible signal at a receiver. It merely carries the intelligence to the receiver. In other words, it is the *carrier*. At the transmitter, the intelligence is impressed on the carrier by a process called *modulation*. When the modulated carrier is picked up by the receiver, the reverse of modulation occurs. At the receiver, the intelligence is removed from the carrier. This process is called *detection*. This chapter gives you the basic principles of modulation and detection, and explains the operation of a large number of detector circuits commonly employed in radar and communications equipment.

## THEORY OF MODULATION

Various methods are used for impressing the signal voltage (intelligence) on the r-f voltage (the carrier) generated by the transmitter. Sometimes, this process (modulation) is accomplished by varying the *amplitude* of the carrier in accordance with the signal voltage. Sometimes, the *frequency* of the carrier is varied in accordance with the signal voltage, and sometimes it is the *phase* of the carrier that is varied. Thus, there are three common types of modulation — amplitude, frequency, and phase.

### Amplitude Modulation

The definition of the popular concept of amplitude modulation is *the process by which the amplitude of an r-f carrier wave is varied at the audio rate of the modulating signal*. This defini-

tion has been accepted for many years because it easily describes the composite AM waveform as viewed on an oscilloscope. But what appears to be an r-f carrier varying at an audio rate is actually the instantaneous frequency-voltage relationship of the output frequency components of a nonlinear device.

Amplitude modulation is more accurately defined as *a type of modulation that results when a modulating voltage is applied to a carrier voltage. It produces amplitude and frequency change of the r-f sidebands*. The carrier itself does not change in amplitude or frequency during the process of amplitude modulation. The radio-frequency carrier is defined as *a single radio frequency radiated by a transmitter operating without modulation*. Since the purpose of this manual is to provide the technician with a basic coverage of the subject, a thorough understanding of the technical terms used is required.

The typical amplitude-modulated carrier is illustrated in Figure 9-1B. You can see that the amplitude of an amplitude-modulated wave is high during the positive peaks of the signal voltage, but is near zero during the negative peaks. The shape of this wave is the result of applying the carrier and signal voltages to a nonlinear device. A nonlinear device is one in which the current does *not* vary in direct proportion to the voltage.

Whenever two signals are impressed on a linear device (one in which the current *does* vary in direct proportion to the voltage), the output waveshape contains the same frequencies as the input and looks like the waveshape in Figure



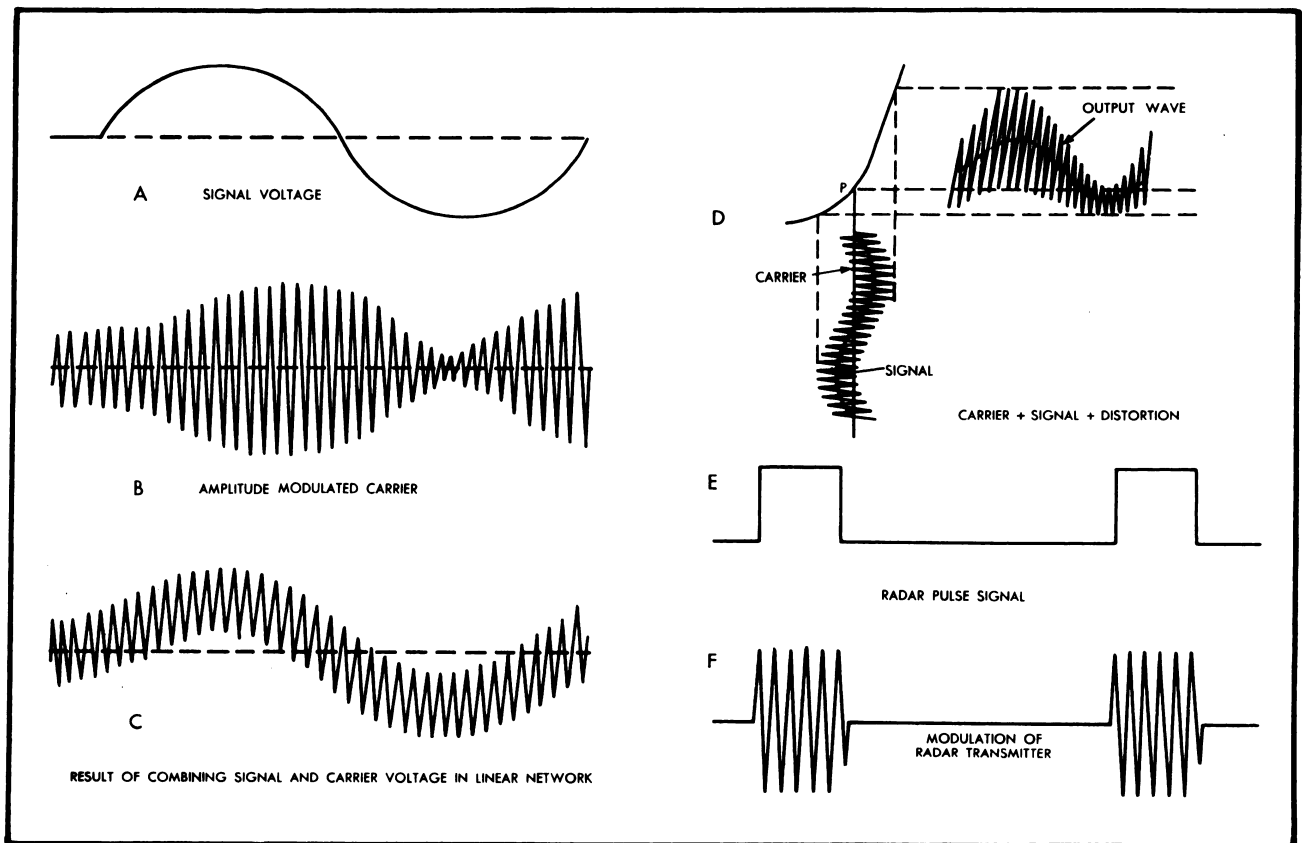


Figure 9-1. Amplitude Modulation

9-1C. In this case the r-f amplifiers in the transmitter amplify the carrier, but discriminate sharply against the modulating frequency. This causes only the carrier to be transmitted. Such an arrangement would defeat the purpose of modulation, as the signal arriving at the receiver would not contain a modulation component at all.

Impressing the carrier and the modulating signal frequencies on a nonlinear network, however, produces the output waveshape shown in Figure 9-1D. The network used in producing this waveshape is a vacuum tube biased to operate on the nonlinear portion of its characteristic curve (point P). The input signal at D is the result of simultaneously impressing the modulating signal and the carrier signal onto the grid circuit of this nonlinear network. The instantaneous grid voltage varies around the bias value as indicated by the input signal. The distorted plate voltage current produces the distorted output waveshape. This is the waveshape you would see on an oscilloscope.

The output of a wave impressed on a nonlinear network contains both the carrier and the signal frequencies and direct current. As you can see in the output waveshape in Figure 9-1D, the negative half-cycles are altered differently than the positive half-cycles. A sine wave altered in this manner is said to be a combination of the fundamental frequencies and additional frequencies caused by distortion. The distortion is introduced by the nonlinear resistance.

When a carrier frequency and a signal frequency are applied to a nonlinear device, the output wave contains the following frequencies:

1. The carrier frequency.
2. The *upper side frequency*, which is equal to the carrier *plus* the signal frequency.
3. The *lower side frequency*, which is equal to the carrier *minus* the signal frequency.
4. Direct current.

When an output wave containing these frequencies is present in a transmitter, the tuned circuits respond to the carrier, the upper side frequency, and the lower side frequency, but not

to the modulating signal. The result of this frequency response is a composite waveshape, like the amplitude-modulated carrier shown in Figure 9-1B. The direct current signal indicates the presence of modulation by varying the plate current meter in the modulator stage of the transmitter.

When an 870-kilocycle carrier frequency of constant amplitude is modulated by a 5000-cps signal, the frequencies in the output are the carrier-frequency of 870 kc, an upper sideband frequency of 875 kc, and a lower sideband frequency of 865 kc. Of these, only the 865 kc, 870 kc, and 875 kc components will be accepted by the output tuned circuits of the transmitter. Therefore, they are the ones that are transmitted. The bandwidth of a signal is measured from the lowest to the highest frequency in the sidebands. In this case it is 10 kc (865 to 875 kc).

An important point to remember in amplitude modulation is that the transmitted signal, which the receiver picks up, does *not* contain the modulating frequency at all. It contains three separated, but related, frequency components. Because of this, receivers require a nonlinear device such as a detector or demodulator to *generate*, or reproduce, the signal frequency intelligence in its original form when the carrier and side frequencies are impressed upon it.

For radar systems, the same principles hold true. A typical signal frequency, such as shown in Figure 9-1E, is a 1-microsecond squarewave pulse which recurs at about 1000-microsecond intervals. This wave is composed of many harmonic (sine wave) frequencies. When a carrier is modulated by these pulses, waveshapes like those at F result which contain the carrier and a wide band of side frequencies. The side frequencies are located on each side of the carrier and are called *sidebands*.

### Frequency Modulation

In frequency modulation the signal is impressed on the carrier by varying the frequency of the carrier in accordance with the frequency of the signal (Figure 9-2). The frequency of the carrier when it is not being modulated is called the *center frequency*. When the carrier is modulated by a positive signal voltage, its frequency becomes higher in proportion to the amount of positive signal voltage, and when it is modulated by a negative signal voltage, its fre-

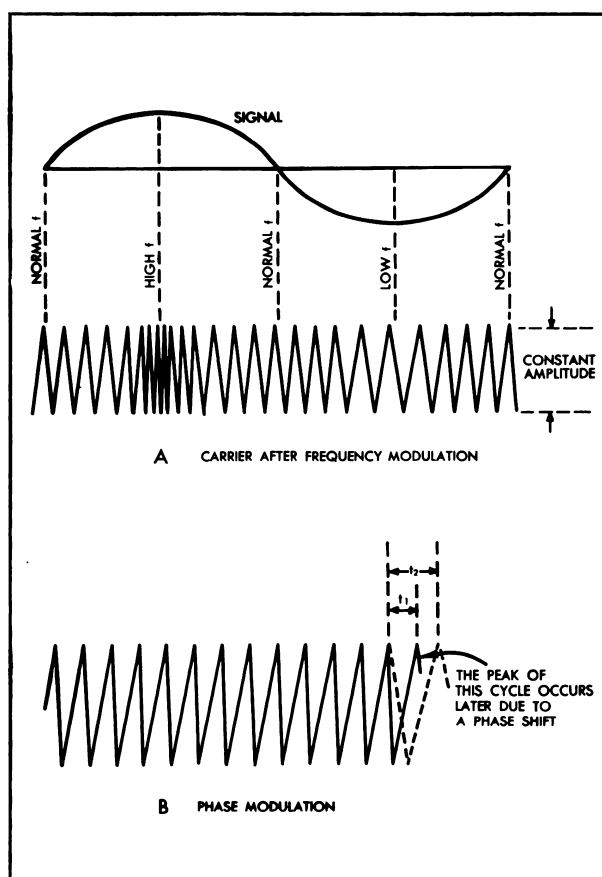


Figure 9-2. Frequency and Phase Modulation

quency becomes lower in proportion to the amplitude of the negative signal voltage. Therefore, when the sine wave shown in Figure 9-2A is applied to a carrier, the carrier frequency changes from the normal center frequency to a higher frequency, back to the normal, then to a lower frequency, and back to the normal frequency again in accordance with the voltage of the applied sine wave. The maximum frequency change from the center frequency, which depends upon the amplitude of the signal, is called the *deviation*. (Note that the amplitude of the modulated carrier is constant.)

One advantage of frequency modulation over amplitude modulation is in its noise reducing capabilities. If the receiver is responsive only to changes in frequency, frequency modulation makes possible a considerable increase in the signal-to-noise ratio. Most noise signals produce amplitude modulation of the carrier, or the carrier-plus-modulation signal, which is applied to the demodulating circuit in the receiver.

Another advantage of frequency modulation over amplitude modulation is that there is no power variation with the degree of modulation, as there is with an amplitude-modulated wave. To obtain 100% modulation with an amplitude-modulated transmitter, the average power output must be increased by 50%. A frequency-modulated transmitter requires an insignificant amount of additional power to produce modulation; thus it needs no provision for increased power output on modulation peaks.

The principle of frequency modulation was developed as early as 1920, but was not put to practical use until the 1940's. Radar equipment uses a special circuit, called a *discriminator*, for detecting frequency-modulated radio signals. This circuit gets its name because it discriminates between different frequencies. The discriminator is taken up later in this chapter.

### Phase Modulation

Phase modulation, which is a slight modification of frequency modulation, is characterized by changes in the phase of the AC carrier voltage instead of in its frequency. This means that the phase of a cycle shifts with respect to each preceding cycle and that the last cycle is ahead of or behind its normal phase. This causes the peak to occur earlier or later than it would normally occur, as you can see in Figure 9-2B, which shows the normal period as  $t_1$  and the longer period as  $t_2$ . The shorter period corresponds to the higher frequency. A frequency modulation detector responds to phase changes in the same manner as it responds to frequency changes.

## REACTANCE TUBE MODULATOR

Reactance tubes serve the function of introducing reactance into associated networks. If the associated network is an ordinary oscillator whose frequency is not stabilized, then the reflected reactance will change the frequency of the oscillations that are generated. Therefore, when the reactance swings or varies, frequency modulation is produced. Considerable frequency deviation can be produced by this method.

Figure 9-3A shows a reactance tube, sometimes called a quadrature tube, which, with its load  $R_1C_1$ , causes reactance fluctuations in the oscillator's frequency-determining LC branch.

From the equivalent network of Figure 9-3B, note that a reactance  $X + \Delta X \sin \omega T$  is generally introduced across terminals 1 and 2 of the frequency determining network of the oscillator. The portion  $X$  represents the steady or fixed reactance in the absence of a modulating signal. The portion  $\Delta X \sin \omega T$  represents the variable reactance component that is introduced by the reactance tube when the modulating signal is applied to it. The variable component changes according to the modulating signal and brings about frequency modulation. The fixed component  $X$  helps stabilize the center frequency of the oscillator.

In Figure 9-3A, capacitors  $C_2$ ,  $C_3$ , and  $C_4$  are coupling elements, while  $C_5$  and  $C_6$  are bypass capacitors. They can have, therefore, no effect on the frequency when their sizes are properly chosen. Capacitor  $C_4$  acts as a feedback coupling capacitor to the oscillator grid. Capacitor  $C_2$  acts as a coupling capacitor to the grid of  $V_1$ .

### Effect With No Modulation Signal

The first effect to consider is that of the fixed value of reactance,  $X$ , on the frequency of the oscillator. This occurs when the modulator tube is connected across the tank circuit with no modulating signal applied. There is a voltage,  $E$ , across the tank circuit ( $L$  and  $C$ ) which is connected between points 1 and 2.  $R_1$  and  $C_1$  in series are also connected between points 1 and 2; therefore, the same voltage appears across them; they are in parallel with the tank circuit. This voltage causes a current ( $I_1$ ) to flow through  $R_1$  and  $C_1$ . If  $R_1$  is at least five times as large as the value of the reactance of  $C_1$ , this branch of the circuit is essentially resistive. The current  $I_1$ , then, can be considered as being in phase with the driving voltage  $E$ . Therefore, the vectors representing  $I_1$  and  $E$  in Figure 9-3C are shown in the same direction (vertical).  $I_1$  flowing through  $C_1$  produces voltage drop  $E_1$  across  $C_1$ .

Across the capacitor  $E_1$  lags  $I_1$  by  $90^\circ$ , as illustrated by their vectors in Figure 9-3C. The voltage  $E_1$  is applied to the control grid of the modulator tube  $V_1$ , producing a corresponding plate current variation,  $I_p$ , through  $V_1$ . The plate current variations of  $V_1$  essentially flow only through the LC tank circuit back toward the cathode of  $V_1$ , since the value of  $R_1$  is high in comparison to the multiple impedance across points 1 and 2. Theoretically, the impedance



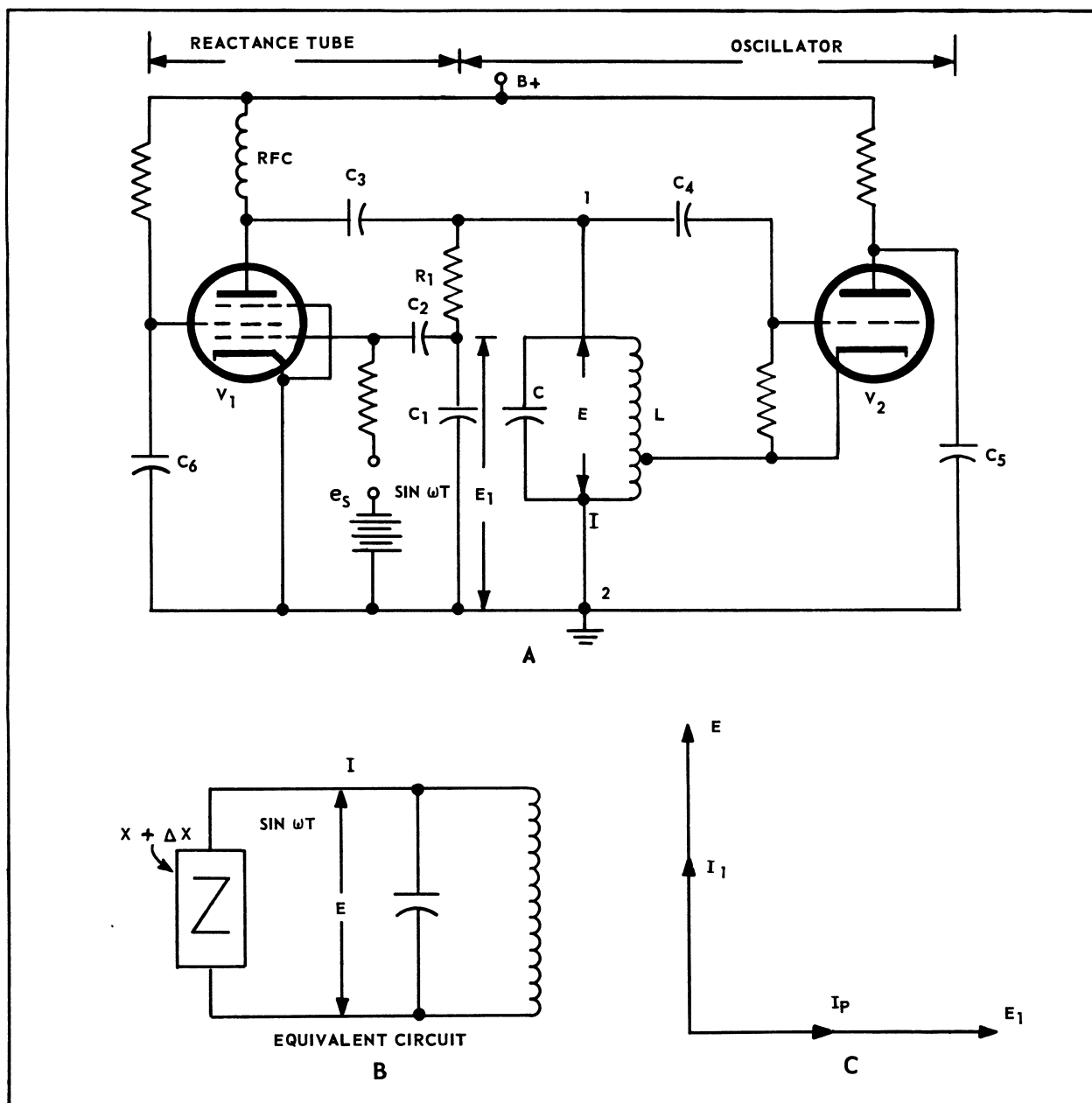


Figure 9-3. Reactance Tube Modulator

across the parallel LC circuit is infinite at resonance if there is no resistance in the circuit. However, in practice this parallel combination of L and C has a much smaller impedance than that offered by  $R_1C_1$  branch. Since current is inversely proportional to impedance, most of  $I_p$  flows through the tank circuit.

As explained previously,  $E$  is the voltage across the tank circuit. The original tank current, labeled  $I$ , is in phase with  $E$  since the tank

circuit is resonant.  $E_1$  lags  $E$  by  $90^\circ$ ; the plate current  $I_p$  is in phase with the grid voltage  $E_1$ . In the LC circuit, then,  $I_p$  is a superimposed current that lags the original tank current ( $I$ ) by  $90^\circ$ . (This  $90^\circ$  out-of-phase action is why  $V_1$  is sometimes called a quadrature tube.)

$I_2$ , the resultant current due to  $I_p$  and  $I$  (Figure 9-4) lags  $I$  and  $E$  by some angle depending upon the relative amplitudes of  $I$  and  $I_p$ . Since  $I_2$  is a lagging current, the impedance across the

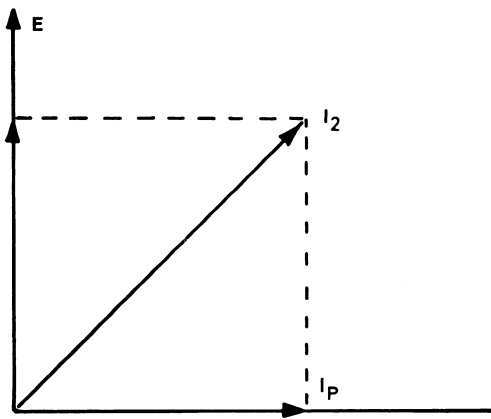


Figure 9-4. Effect of  $I_p$

tank circuit cannot be at its maximum unless something happens. What does happen is that the frequency of oscillation of the tank changes until the voltages across the tank and the current flowing into it are again in phase. At this time the impedance across the tank is again maximum. This is the same result as if a reactance had been introduced in parallel with the LC tank circuit. Since the current  $I_p$  lags the voltage  $E$  by  $90^\circ$ , the introduced reactance is inductive. Earlier you learned that inductance decreases as inductors are added in parallel. Thus, in this case the frequency of the oscillator increases to a new fixed value due to the introduced reactance  $X$ .

If the components  $R_1$  and  $C_1$  (Figure 9-3A) were reversed in position, the grid voltage  $E_1$  would be that across  $R_1$ . In this case, the reactance of  $C_1$  should be at least five times the value of  $R_1$ , so the voltage  $E_1$  will lead  $E$  by

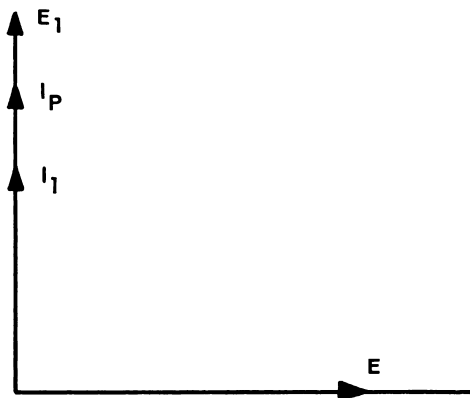


Figure 9-5. Effect of Reversing  $R_1$  and  $C_1$

approximately  $90^\circ$ , as shown in Figure 9-5. The plate current  $I_p$  will then lead  $E$  by  $90^\circ$ . The introduced reactance  $X$  has a capacitive effect on the oscillating circuit, just as if a capacitor had been connected in parallel with it. A decrease in frequency occurs.

### Summary

The modulator tube is connected across an oscillator tank circuit. The modulator tube is made to appear as either a capacitance or an inductance by having its grid excited with a voltage which either leads or lags the oscillator voltage by  $90^\circ$ .

A leading grid voltage causes a corresponding leading plate current which makes the reactance tube appear as a capacitor in parallel with the oscillator tank circuit. This results in a decrease in the center frequency of the oscillator. A lagging grid voltage causes a corresponding lagging plate current which makes the reactance tube appear as an inductance in parallel with the oscillator tank circuit. This results in an increase in the center frequency of the oscillator. A reactance tube which acts as an inductance can be made to act as a capacitance only by making the proper circuit changes so the grid voltage leads the oscillator tank voltage.

### Effect With Modulating Signal

The effect of connecting the modulator tube across the tank circuit when no modulating signal is applied has been shown. Now see what happens when a modulating signal is applied. The magnitude of the introduced reactance is determined by the magnitude of  $I_p$  for a given amplitude of grid voltage,  $E_1$ . The magnitude of  $I_p$  for a given  $E_1$  is determined by the transconductance of  $V_1$  (Figure 9-3). That is,  $g_m = I_p/E_1$ .

Therefore, it may be said that the value of reactance introduced into the tuned circuit varies directly with the  $g_m$  of the reactance tube. When the modulating signal is applied, both  $E_1$  and  $I_p$  change, thus causing  $g_m$  to vary with the modulating signal. At this time a variable reactance,  $\Delta X \sin \omega T$ , is introduced into the tuned circuit.  $\omega$  in this variable reactance either adds to, or subtracts from, the fixed value of reactance that is introduced in the absence of the modulating signal. It is this variable reactance that

changes the frequency of the oscillator in accordance with the modulating signal. When properly designed and operated, the reactance tube modulator gives linear frequency modulation, and produces great frequency changes in accordance with the modulating signal.

Only one reactance tube modulator circuit has been presented. There are various types of reactance tube circuits that will produce frequency modulation. The main difference in the various arrangements is the type of phase-shifting circuit used to give a grid voltage that is  $90^\circ$  out of phase with the r-f voltage of the oscillator. All reactance tube modulators produce the same thing — linear deviation over quite a wide frequency band.

### DETECTION

Previously it was stated that an amplitude-modulated waveshape contains the radio frequencies, but not the signal frequencies, and that it is the function of the detector in the receiver to reproduce the signal frequencies from the variation of the carrier and its sidebands.

Detection of amplitude-modulated signals requires a nonlinear electrical network. An ideal nonlinear curve for this is one that affects the positive half-cycles of the modulated wave differently from the negative half-cycles, and that so distorts a wave with an applied voltage of zero average value that the average resultant current varies as the signal frequency. The curve shown in Figure 9-6A is called an ideal curve because it is linear on each side of the operating point P and does not introduce harmonic frequencies.

When the input to an ideal nonlinear curve is a carrier and its sidebands, the output contains the following frequencies:

1. The carrier frequency.
2. The upper sideband.
3. The lower sideband.
4. A frequency equal to the carrier minus the lower sideband (or, a frequency equal to the upper sideband minus the carrier), which is the original *signal frequency*.

The detector reproduces the signal frequency by producing a distortion of a desirable kind in its output circuit. When the output voltage of the detector is impressed upon a low-pass filter,

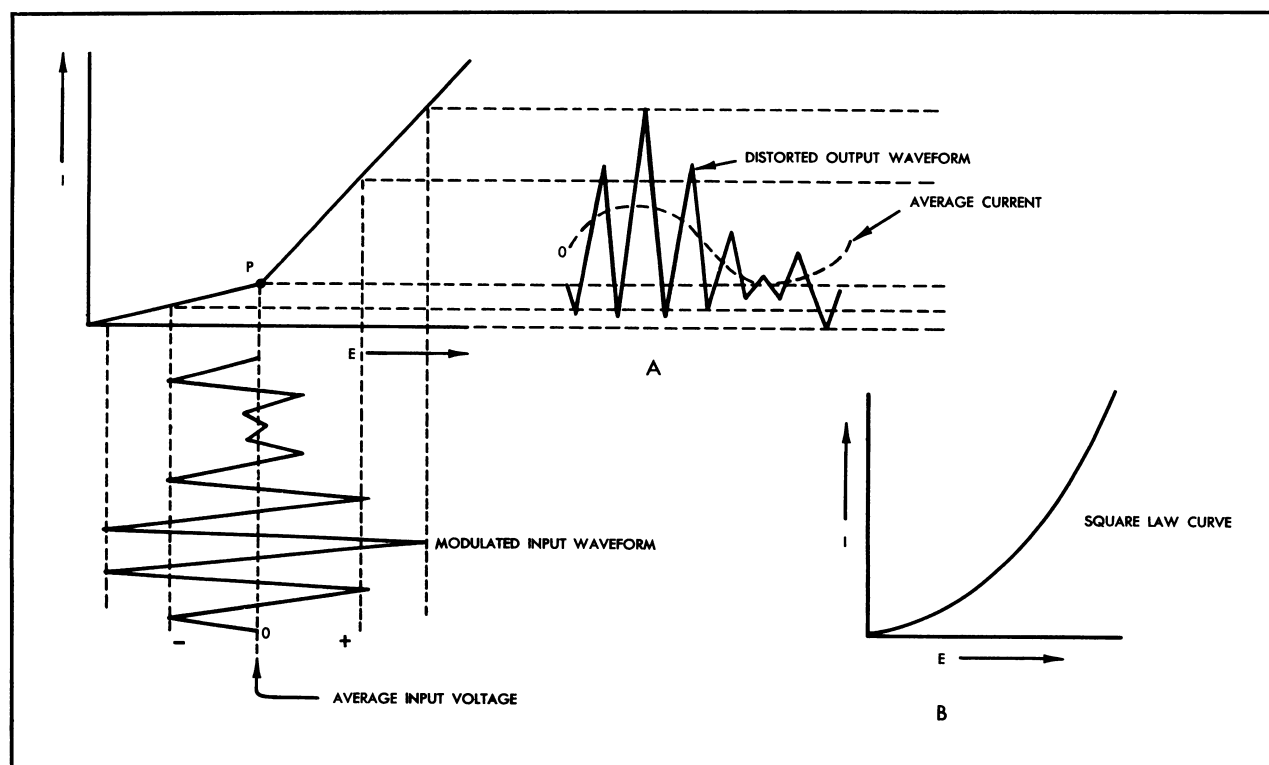


Figure 9-6. Nonlinear Device Used as Detector

which suppresses the radio frequencies, only the original or signal frequency is left.

In some practical detector circuits, the nearest approach to the ideal curve is the square-law curve shown in Figure 9-6B. The output of a device using this curve contains, in addition to all the frequencies which were listed, the harmonics of each of the input frequencies. The harmonics occur because voltage inputs that have a large amplitude are distorted differently from voltage inputs that have a low amplitude. The harmonics of radio frequencies can be filtered out, but harmonics of the signal frequencies, even though they produce an undesirable distortion, have to be tolerated. However, the square-law curve offers an advantage in that the output amplitude varies as the square of the input. In contrast, a linear curve produces an output which varies directly with the input. Therefore, the square-law device produces a greater output for a given input.

### Diode Detector

One of the simplest and most effective types of detectors and one with nearly an ideal non-linear resistance characteristic is the diode detector. Notice the  $E_p-I_p$  curve in Figure 9-7B. This is the type of curve on which the diode detector at A operates. The curved part of the curve is the region of low plate current, and indicates that for small signals the output of the detector will follow the square law. For input signals with large amplitudes, however, the output is essentially linear in the positive direction from the operating point. This type of detector is classed as a power detector since it handles large input amplitudes without much distortion.

The modulated carrier is introduced into the tuned circuit made up of  $LC_1$ . This circuit is designed so that the receiver has high selectivity. The waveshape of the input to the diode plate is shown in Figure 9-7C. As a diode conducts only during positive half-cycles, this circuit removes all the negative half-cycles and gives the result shown in Figure 9-7D. The average output is shown at E. Although the average input voltage is zero, the average output voltage across R always varies above zero and has an average voltage of  $\frac{1}{2} \times 0.637 \times$  peak voltage for any positive half-cycle.

The low-pass filter, made up of capacitor  $C_2$  and resistor R, removes the r-f (carrier fre-

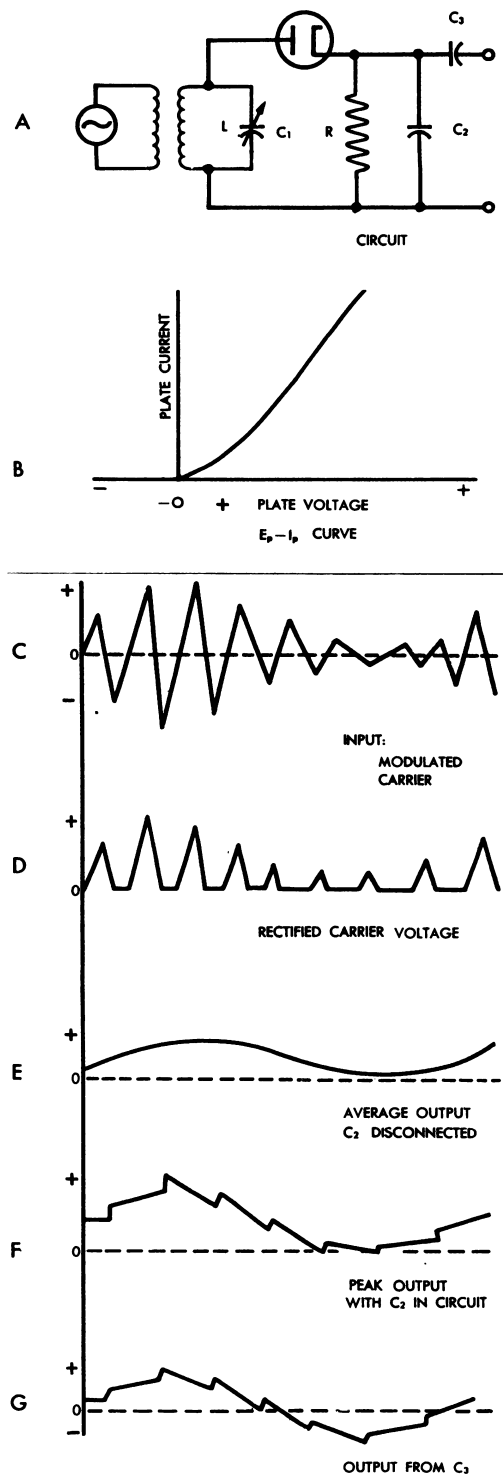


Figure 9-7. Diode Detector

quency) which, so far as the receiver is concerned, serves no useful purpose. Capacitor  $C_2$  charges rapidly to the peak voltage through the



small resistance of the conducting diode, but discharges slowly through the high resistance of  $R$ . The sizes of  $R$  and  $C_2$  normally form a rather short time constant at the signal (audio) frequency and a very long time constant at the radio frequencies. The resultant output with  $C_2$  in the circuit is a varying voltage that follows the peak variation of the modulated carrier (see Figure 9-7F). The DC component produced by the detector circuit is still in the waveshape but may be removed by capacitor  $C_3$ , producing the AC voltage waveshape in Figure 9-7G. In communications receivers the DC component is often used for providing automatic volume (gain) control.

Advantages of diode detectors are as follows:

1. Ability to handle relatively high-power signals. There is no practical limit to the amplitude of the input signal.
2. Low distortion. Distortion decreases as the amplitude increases.
3. High efficiency. When properly designed, 90% efficiency is obtainable.

4. Diode detectors develop a readily usable DC voltage for the automatic gain control circuits.

Disadvantages of the diode detectors are:

1. Power is absorbed from the tuned circuit by the diode circuit. This reduces the  $Q$  and selectivity of the tuned input circuit.
2. No amplification occurs in a diode detector circuit.

Careful selection of component parts is necessary for obtaining optimum efficiency in diode detector circuits. One very important fact to consider is the value of the time constant  $RC_2$ , particularly in the case of pulse modulation. When a carrier modulated by a square pulse (Figure 9-8B) is applied to an ideal diode detector, the waveshape shown in Figure 9-8C is produced. Notice that for clarity the amplitude of the wave at C is exaggerated in comparison to the high-frequency carrier shown at B.

If the time constant of  $RC_2$  is too long, several cycles are required to charge  $C_2$ , and the leading edge of the output pulse is sloped as shown in Figure 9-8D. After the pulse passes by, the capacitor charges slowly and the trailing edge is exponential rather than square as desired. If the time constant is too short, both the leading and trailing edges are steep. However, the capacitor discharges considerably between cycles, and this reduces the average amplitude of the

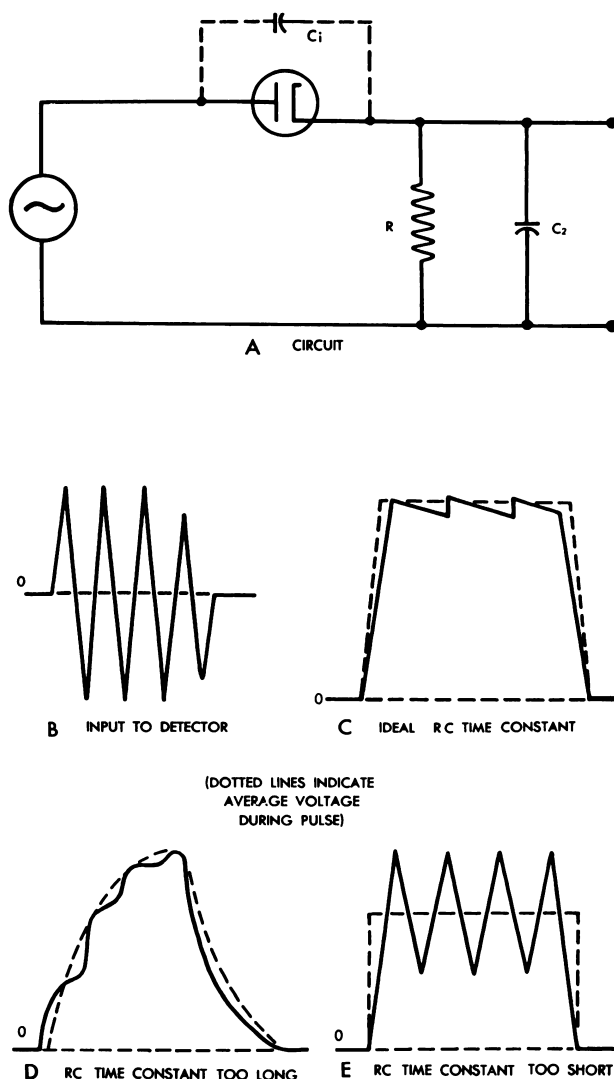


Figure 9-8. Diode Detector Component Considerations

pulse, leaving a sizable component of the carrier frequency in the output as shown in Figure 9-8E.

For these reasons the selection of the time constant is a compromise between an output that is large but distorted, and an output that is undistorted but small and only partly filtered. To achieve this compromise, the load resistor  $R$  must be large, as the total input voltage is divided across  $R$  and the internal resistance of the diode when the tube is conducting. A large value of load resistance insures that the greater part of this voltage will be in the output where it is desired. On the other hand the load resistance must not be so high that capacitor  $C_2$  becomes

small enough to approximate the size of  $C_1$ , the internal capacitance of the tube. When this occurs, capacitor  $C_2$  will try to discharge through  $C_1$  during the nonconducting periods, which would reduce the amplitude of the detector output.

### Grid-Leak Detector

The grid-leak power detector is similar to the diode detector with the advantage that it serves as a stage of amplification. Grid-leak detectors use no fixed bias arrangement in the grid circuit; they depend for their operation on the flow of grid current whenever the grid is positive.

In the typical grid-leak circuit illustrated in Figure 9-9A, the grid, cathode,  $C_4$ ,  $R_1$ , and the tuned circuit form a diode detector circuit. If capacitor  $C_4$  and resistor  $R_1$  are correctly proportioned, the capacitor charges during the period of current flow and discharges slowly through  $R_1$  during periods of no current flow. Since the capacitor maintains an average charge, it places an average negative voltage on the grid with respect to the cathode. (This voltage is represented by the heavy line in the grid voltage curve in Figure 9-9D.)

The r-f voltage at the grid varies around this average value. If the capacitor should discharge

slightly, the next positive cycle of the modulated voltage will go higher in the positive direction, recharging the capacitor. As the modulated waveform decreases in amplitude (that is, as the positive peaks become smaller), the capacitor does not fully recharge with each cycle, and the average charge (grid bias) reaches a lower value.

This means that the average grid voltage varies in accordance with the variation of the amplitude of the modulated wave and, in doing so, reproduces the signal voltage. Consequently, the plate current and the plate voltage shown in Figure 9-9E also vary at the signal rate. The low-pass filter in the plate circuit prevents the plate voltage from changing at an r-f rate and thus removes the radio frequencies from the output. The normal gain of the triode stage produces an amplified waveshape in the plate circuit. The time constant of  $C_4R_1$  is governed by the same factors as the filter used in the diode detector.

The voltage applied to a grid-leak detector must not be so high that it causes the average grid voltage to exceed the plate current cutoff voltage for the tube. This characteristic limits the power handling capacity of the grid-leak detector. Like the diode detector, the grid-leak detector produces little distortion when the received signal is strong. Although it produces

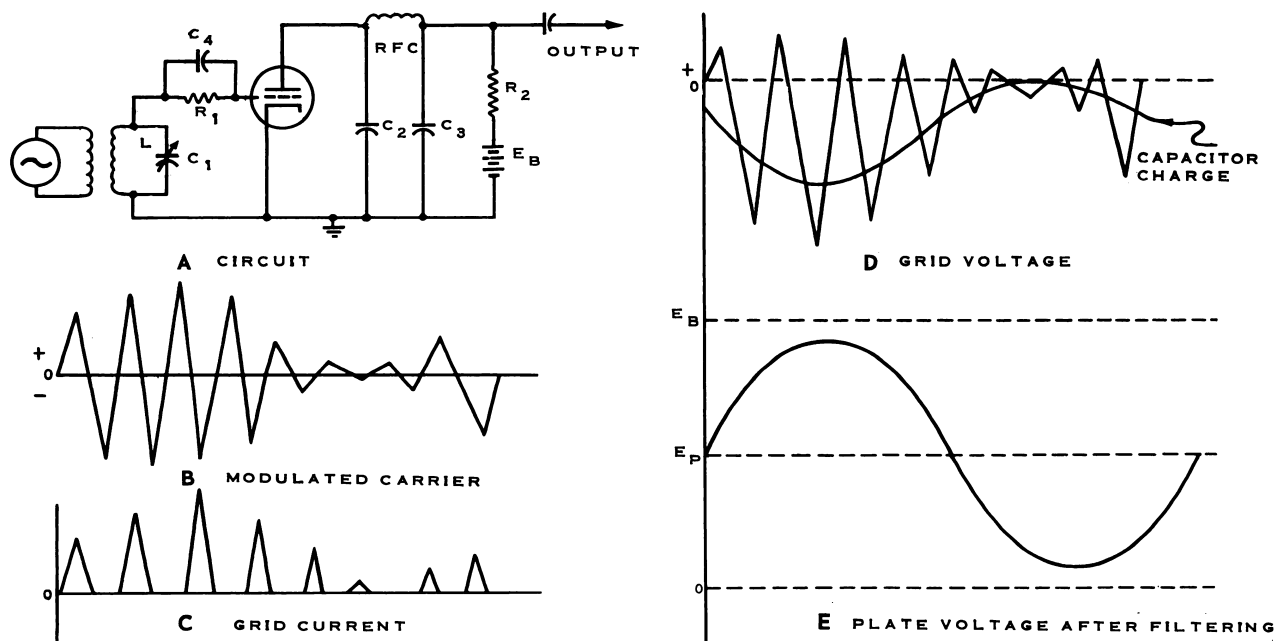


Figure 9-9. Grid-Leak Detector

only a limited amount of DC voltage for AVC, the grid current required for this purpose reduces the Q of the preceding stage. Due to the amplification properties of a grid-leak circuit, its output for a given input is much greater than that of the diode detector.

### Plate Detector

The plate detector shown in Figure 9-10B, sometimes called a bias detector or plate detector, is similar to a class A amplifier stage except that it is operated at very near cutoff bias. For ideal operation, this bias may be provided from a fixed DC potential, but for practical purposes, a properly bypassed cathode resistor will bias the tube near enough to cutoff. For this reason  $R_k$ , the cathode resistor, shown in Figure 9-10B, is rather large, and the bypass capacitor  $C_k$  is large enough to bypass both carrier and modulation frequencies. Because of the extremely high operating bias, the tube characteristics are very nonlinear as you can see by examining the  $E_g-I_p$  curve at A. Plate current flows only during positive halves of the input signal and a small portion of the negative halves. The grid does not draw current as long as the signal does not drive the grid into the positive region of the  $E_g-I_p$  curve. The average plate current, shown at A by dashed lines, varies as the modulation varies.

The removal of the r-f component from the plate circuit is accomplished by the low-pass

filter, formed by the choke RFC and capacitor  $C_1$  and  $C_2$ . The variation remaining after removal constitutes the original modulation. Because of the amplification factor of the tube, the voltage representing the modulation which appears across the load resistor is greater than the input (grid) voltage. All other AC voltages are effectively placed at ground potential by capacitor  $C_1$ .

When a pulse-modulated signal is applied to the detector, a pulse appears as a positive pulse at the grid and as a negative pulse in the plate circuit. In some applications positive output pulses are desired. This is done by changing the circuit so that  $R_L$  is located in the cathode-to-grid circuit instead of in the plate circuit. By this arrangement the output has the correct polarity and may be taken across  $R_L$ . The other ground connections are removed, as you can see by examining the inverted output circuit for pulse type modulation in Figure 9-10C.

The advantage of the plate detector is that since the grid does not draw current, the Q of the input circuit remains the same and amplification occurs in the circuit. Its disadvantages are higher distortion and lower power handling ability than the diode detector.

### Infinite-Impedance Detector

The infinite-impedance detector is a modified type of plate detector. It has the same ability

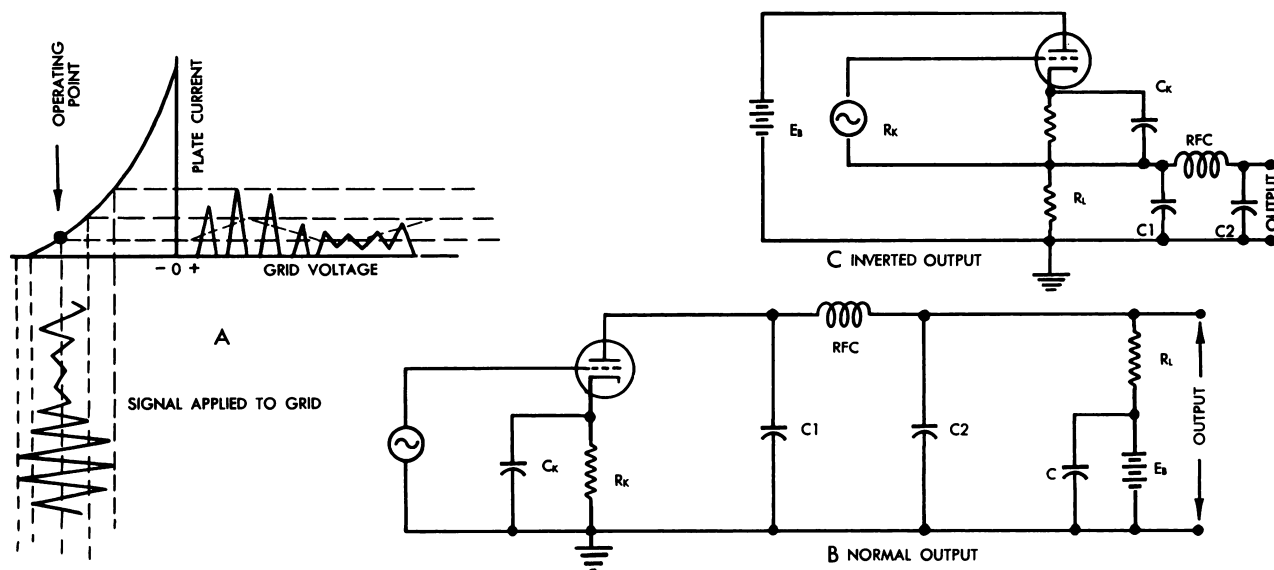


Figure 9-10. Plate Detector

as the diode detector to handle large input amplitudes with little distortion but without drawing current from the circuit driving it. Drawing current by any type of detector causes reduction in the  $Q$  of the tuned circuit and a decrease in the selectivity.

In Figure 9-11, which shows the infinite impedance detector,  $E$  is the modulated carrier voltage.  $LC_1$  is a tuned circuit that contributes to the selectivity of the receiver. The plate of the triode is connected directly to the power supply. Capacitor  $C_3$  is a virtual short circuit to ground for AC voltages. The load resistance  $R_1$  is in series with the cathode. Capacitor  $C_2$ , in conjunction with  $R_1$ , forms a long RC to the carrier frequency but a short RC to signal frequencies.

Under no-signal conditions the plate current through  $R_1$  provides a bias that keeps the grid very near cutoff. Under signal conditions when an r-f voltage is applied, the positive peak of the modulated voltage charges capacitor  $C_2$  to a higher voltage. This capacitor holds its charge between peaks because of the long time constant of  $R_1C_2$ . No grid current flows during the positive peaks because the cathode bias changes at the same time that grid voltage changes, and the cathode-grid voltage difference is never great enough to exceed the cutoff value. When the crest of the modulation peak has passed, the amplitude of the peak decreases and the capacitor discharges at the same rate. Therefore, the capacitor charge follows the modulation peaks in the same manner as in the diode detector.

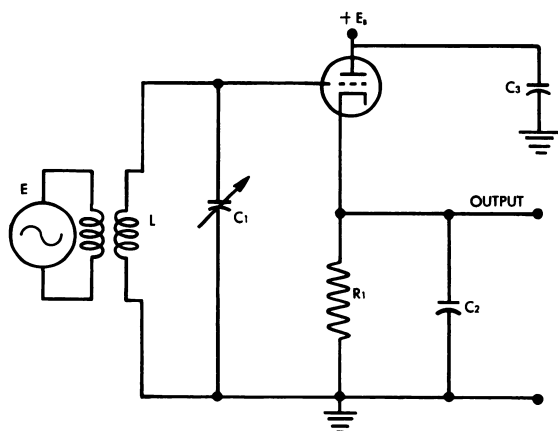


Figure 9-11. Infinite Impedance Detector

The infinite-impedance detector employs a very high input impedance for two reasons. First, the grid does not draw current. Second, because of the capacitance in the plate circuit, the impedance reflected into the grid circuit is purely resistive.

Another characteristic of this detector is that, like the cathode follower, negative feedback exists because of the common input and output impedance at signal frequencies. This negative feedback is responsible for reducing the distortion due to characteristics of the vacuum tube.

Distortion is also minimized by the employment of high input voltages. This input voltage can be as large as half the DC voltage on the plate without overdriving the tube. The ability of the tube to handle large signals is due largely to the fact that the bias increases as the signal input increases.

### Crystal Detector

A small piece of silicon, galena, iron pyrite or carborundum is capable of providing the nonlinear characteristic required for detection. When a crystal of one of these elements is in contact with a sharp pointed wire, the arrangement offers a different resistance to current flow from the crystal to the wire than it does from the wire to the crystal. Such a device is called a crystal rectifier or a crystal detector. (Do not confuse this type of crystal with the piezoelectric crystal used to control the frequency of oscillators.)

For use in radar sets, the crystal is silicon and the wire is tungsten. Figure 9-12D shows the construction. Since some spots on a crystal are more sensitive than others, the point of the tungsten wire is placed in contact with the most sensitive part, and the contact is made permanent by using a sealing compound and mounting the assembly in a watertight, standard-sized cartridge.

Crystals are most frequently used in the mixer stage in the superheterodyne receivers of radar sets. Seldom are they used as detectors to demodulate the modulated carrier. Their use as mixers requires the same nonlinear characteristics as are required in a detector circuit.

Mixer crystals are designated according to use and frequency. The crystal suitable for 10-cm waves comes in at least three types, as does the crystal for 3-cm waves. Three types for each

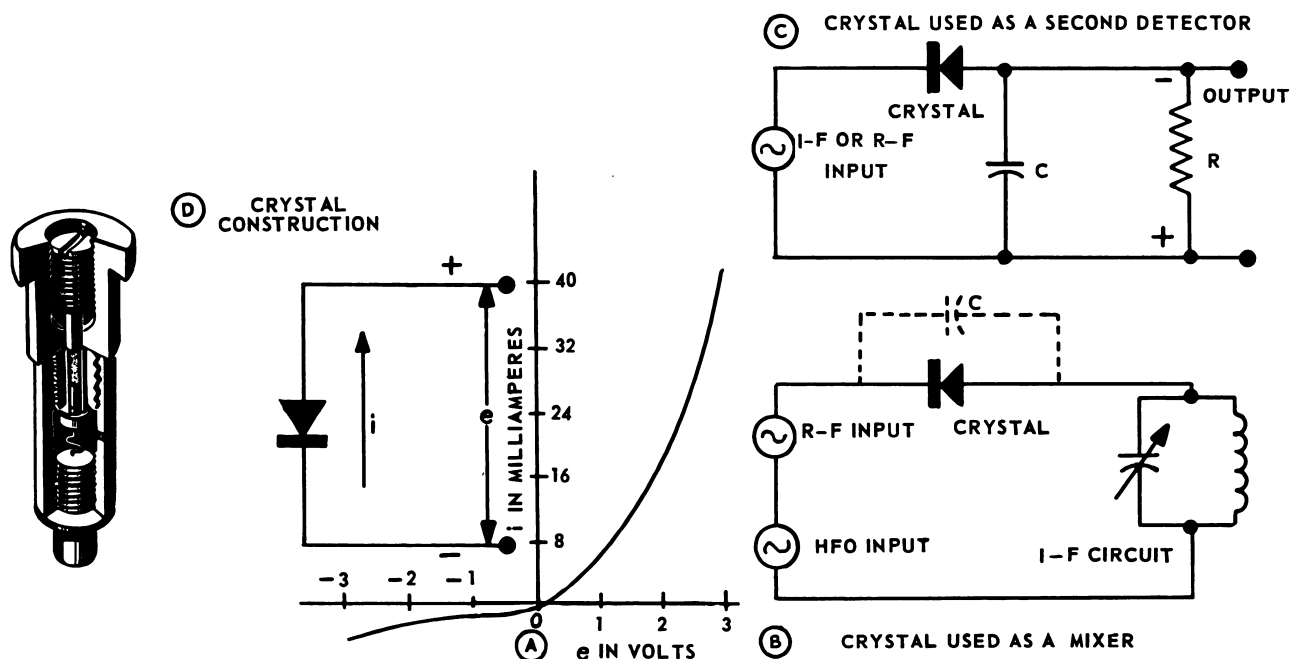


Figure 9-12. Crystal Detector

frequency are characterized by poor, medium, and high-frequency sensitivity. Another type of crystal is used for detector circuits.

As shown in the crystal mixer circuit in Figure 9-12B, the radio-frequency and the local oscillator signals are applied to the resonant circuit through the crystal. The nonlinear characteristics of the crystal cause the voltage across the resonant circuit to contain the sum and difference frequencies and, in addition, the original frequencies. The coil-capacitor circuit resonates to the difference frequency. The other frequencies are rejected and the difference frequency is applied to the i-f amplifiers.

The second detector circuit in Figure 9-12C using the crystal operates similarly in all respects to the diode detector.

Notice in the DC voltage-current characteristic curve of a crystal in Figure 9-12A that the current flow in the crystal, caused by negative voltages, is very small.

You can see from the curve in Figure 9-12A that with two volts negative applied, the current is approximately 2 ma in the reverse direction. Computing the resistance of the crystal we find it to be approximately 1000 ohms. With two volts positive, the curve shows the resistance is about 100 ohms. The back-to-front ratio, or the ratio of the resistance to current flow in a for-

ward direction, to the resistance in the opposite direction, is 10 to 1 for this crystal.

When the crystal is used as a rectifier at 3000 mc, the capacity between the point and the crystal is about 1 mmf, making the capacitive reactance about 50 ohms. This reactance is in parallel with the 1000-ohm DC resistance, making the back impedance about 47 ohms.

Measuring the resistance of a crystal in each direction with an ohmmeter is a rough test of crystal sensitivity. When you know that the back (high) resistance of a crystal has dropped to one-half the original value, the sensitivity of the crystal is too low for that crystal to be used. The only solution is to replace the crystal with a known good crystal.

An ohmmeter test is not conclusive because of the change in back impedance at high frequencies. Since the factory test equipment is not available in the field, you may test the crystal at high frequencies by comparing the signal-to-noise ratio of the receiver with a questionable crystal installed against this ratio with a good crystal installed.

Even though the contact point causes reduced sensitivity, the crystal is superior to vacuum tube mixers at 3000 and 10,000 mc since it introduces less noise into the receiver system. This means a better signal-to-noise ratio. In addition, there



are no transit time difficulties involved in crystals. Disadvantages of crystals are that they damage easily, even while installed in the radar set, and have a gain of less than one. Damage in the radar set usually results from inadequate protection from the transmitter pulse. As a second detector, the crystal has the advantage of being smaller in size than vacuum tube detector circuits and requiring no heater voltage.

### Superregenerative Detector

The superregenerative detector is an extremely sensitive detector. It has a gain comparable to a complete superheterodyne receiver. On examining the circuit of the superregenerative detector in Figure 9-13, you can see that it consists of an r-f oscillator and a tuned circuit that is resonant to the frequency being detected. The superregenerative detector also includes a special circuit that allows the detector to break into oscillation, then quench, or stop the oscillations at regular intervals. This action is accomplished by variations in the detector plate voltage. When the plate voltage is above a certain value, the

circuit oscillates; below this value, it ceases oscillations.

The voltage supplied by the special circuit is called the quench voltage. It varies the plate voltage above and below the value just mentioned at a rate above the audible range as shown in Figure 9-13C. As the plate voltage gets near what is called the critical value, small irregular current changes in the cathode emission cause the plate voltage to rise above the critical point, and the circuit oscillates until the plate voltage drops below the critical value. When the plate voltage comes back up again, the process is repeated. When a radio signal is applied to the tuned circuit, the signal adds enough voltage to the grid circuit to cause oscillations at a somewhat lower value of plate voltage. This causes oscillation to start sooner, and to last until the plate voltage drops back to the critical value.

Under signal conditions, the duration of the oscillation cycle is longer than under no signal conditions. Within limits this time varies directly with the signal strength. Short periods of oscillation build up small charges on the grid leak capacitor, while long periods of oscillation cause

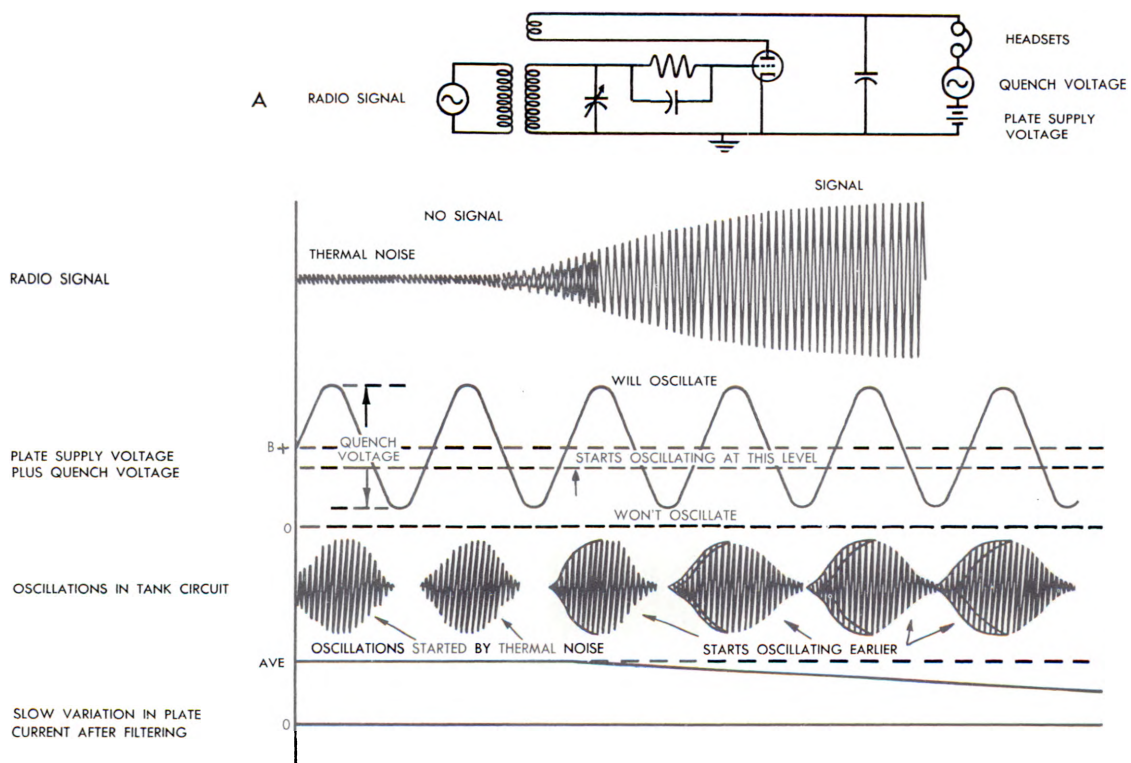


Figure 9-13. Superregenerative Detector



grid current to flow longer and consequently build up greater charges. The charge, in turn, determines the average plate current and it, in turn, varies with the signal strength. If the signal is amplitude-modulated, the variation in plate current will roughly resemble the modulation voltage.

Since the incoming radio signal only initiates the oscillation while the circuit itself determines the amplitude, the circuit produces a tremendous gain when properly adjusted. In fact, any signals stronger than the thermal noise level are audible. Since this is the limit of amplification in a superheterodyne receiver, the sensitivity of a superregenerative detector is near that of the superheterodyne receiver.

The superregenerative detector has the advantage of high gain and light-weight components. It finds its greatest use at very high frequencies where distortions and noise are compensated for by the high gain, which is not obtainable with conventional circuits. Another advantage is that the output voltage is about the same for weak or strong signals. Weak signals bring it to full oscillation equally as well as strong signals do. (For this reason any noise voltages stronger than the signals are reducible to the level of the signal. Suppressing ignition noises and other types of interference further reduces the annoyance to the listener.)

Its disadvantages are excessive distortion, poor selectivity, and a characteristic hiss when no signal is being received. The poor selectivity is because the detector responds to signals off frequency as well as to those on frequency, even though the amplitude of the off-frequency signal is less. The hiss, when no signal is being received, is from the noise produced by the ran-

dom thermal voltages in the tube that start oscillations at irregular times. The hiss disappears when a signal is present, since the signal has a definite amplitude and since the oscillations start at a definite time. The distortion in the detector results from grid current clipping the positive peaks and from accentuation of the negative modulation peaks.

## FREQUENCY CONVERSION

Many stages of amplification are required in a radar receiver to make a very weak echo voltage at the antenna give a good indication on the radar indicator screen. In the simplest type receiver, there are several r-f stages followed by a detector and by a video amplifier. In spite of this, the gain (amplification) and selectivity of an r-f amplifier decreases as the operating frequency increases. Therefore, it is desirable to convert the high radio frequency to a lower radio frequency, and then to amplify the lower frequency. Furthermore, it is easier to design amplifiers with a higher gain when the frequency is constant. The modern receiver employs a high gain and highly selective amplifier in the intermediate-frequency stage, designed to operate at the one frequency to which all received frequencies are converted.

Frequency conversion in a superheterodyne receiver is accomplished by mixing the r-f frequency with the local oscillator frequency (Figure 9-14). This mixing produces a frequency equal to the difference of the two frequencies. If the two frequencies are nearly the same, the difference frequency will be low, but not as low as the audio frequencies. This frequency lies between

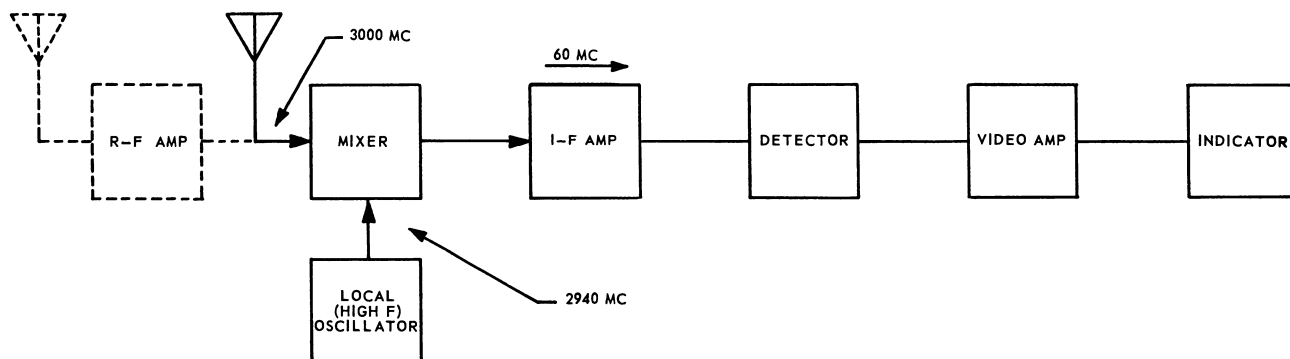


Figure 9-14. Block Diagram of Superheterodyne Receiver

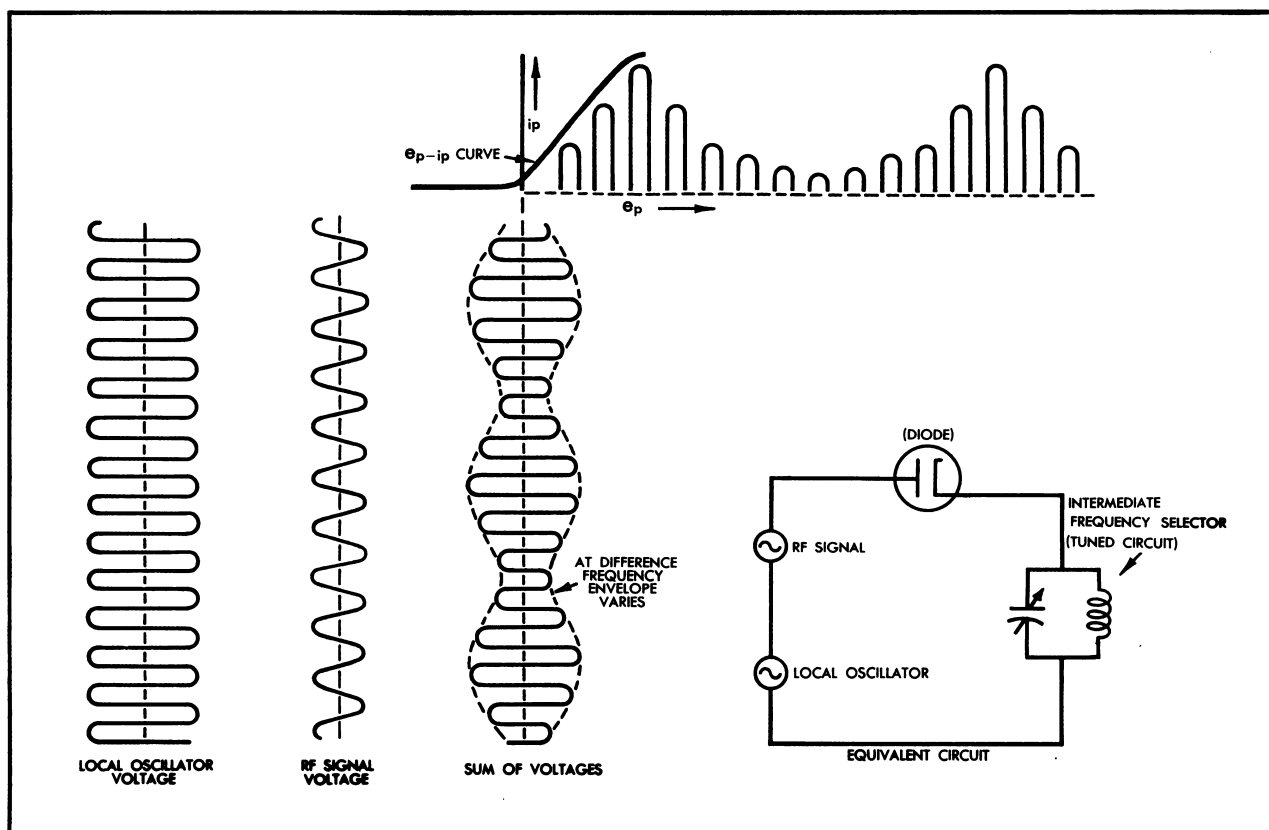


Figure 9-15. Mixer Action

the carrier and signal frequencies, and is called the *intermediate frequency*. If the carrier frequency is below 1000 mc, it is usually amplified by an r-f amplifier before being fed to the mixer. If higher, the r-f signal is introduced directly into the mixer. After mixing, the intermediate frequency is amplified by an i-f amplifier and then is detected in the same manner as any other signal.

Production of a difference frequency in the mixer stage requires the use of a nonlinear electrical circuit for the generation of the new frequency. The mixer, as this circuit is called, produces the new frequency in the same way as a modulator or a detector. For this reason this stage was often called the *first detector*.

When two slightly different frequencies are added, as shown in Figure 9-15, the waveshape representing the combination of the local oscillator frequency and the r-f signal contains only these two frequencies. After distortion by the mixer the waveshape contains not only these two frequencies, but a frequency equal to the sum of the two frequencies and another equal to the

difference between the two frequencies. Since it is the difference frequency that is desired, the tuned circuits in the i-f amplifier select this frequency and the stage amplifies it. If the r-f signal is modulated, the difference frequency (i-f) will also be modulated in the same manner.

Any of the detector circuits previously discussed will work in this stage. Grid-leak and plate detectors are used in the mixer stages of communications receivers in order to utilize the gain of the triode or pentode used. In low-frequency radar sets, the diode is sometimes used. In high-frequency sets, however, the transit time in the tube and the thermal noise it introduces make the vacuum tube inferior to the crystal.

Most 10-cm and all 3-cm sets use the silicon crystal for the nonlinear element in the mixer stage. In the simplified crystal mixer circuit shown in Figure 9-16, currents from both the r-f signal and local oscillator signal flow through the crystal. The difference frequency is selected by the tuned circuits which are connected to the grid of the first amplifier tube. Any nonlinear device produces a unidirectional current that may

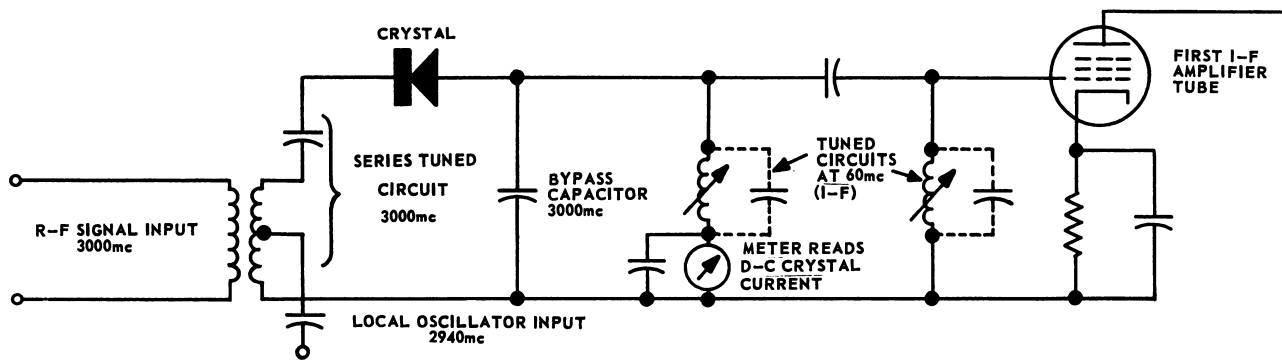


Figure 9-16. Crystal Mixer

be read on a meter. The adjustment of the oscillator for maximum output is determined by observing the meter reading, which varies directly with the amplitude of the oscillator signal.

### DISCRIMINATOR CIRCUIT

Even though the design and construction of radar equipment is precision work, the frequency of a transmitter may shift. This causes considerable trouble at the receiver. Since the transmitter is expected to operate on just one frequency, the local oscillator in the receiver is set on the frequency which, when combined with the input signal (same frequency as the transmitter), will give the desired intermediate frequency. When the transmitter shifts frequency, the difference between this new frequency and the local oscillator frequency will no longer give the preset i-f.

Signals which supply weak returns may no longer be visible on the indicator screen. Any change of the transmitter frequency, therefore, requires either continued readjustment of the receiver tuning by the operator or the use of an automatic tuning device. Constant frequency checking by the operator not only is time consuming, but certainly is not in the interest of highest efficiency. Therefore, most radar receivers have an automatic frequency control circuit in which changes of frequency are detected by a discriminator circuit.

#### Circuit Operation

In a discriminator, a transformer, due to special connections, distorts the frequency change into a voltage of varying amplitude. After distortion, this voltage of varying amplitude is fed

to an amplitude modulation detector. Figure 9-17 shows the circuit of a typical discriminator circuit.

The two coils  $L_1$  and  $L_2$  are the primary and the secondary, respectively, of the i-f transformers. The secondary coil is tuned to the correct transmitter frequency by capacitor  $C$ . Tubes  $V_1$  and  $V_2$  are diode detector tubes. The filters  $R_1C_1$  and  $R_2C_2$  serve the same purpose as the filters in the diode detector, that is, the removal of the r-f component from the circuit.

The presence of  $L_3$  and its connection to the primary at the top and to the secondary at the center presents an unusual feature in the discriminator circuit.  $L_3$  is an r-f choke which has a high reactance to the radio frequency. Since it is connected across the primary, the primary voltage appears across it at all times. The connection to each diode from this choke causes the primary voltage ( $E_p$ ) to appear at the plates of the diode with the same phase shift in each case. This voltage causes currents to flow in opposite directions in resistors  $R_1$  and  $R_2$ , resulting in

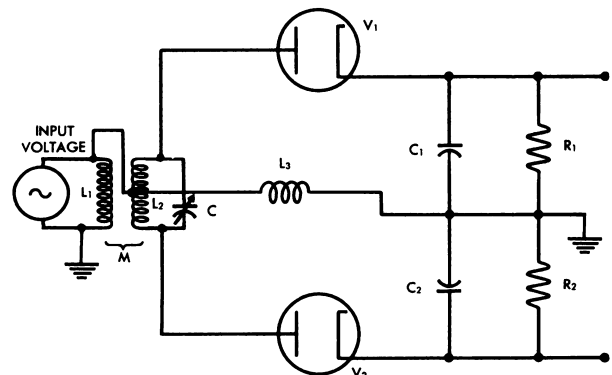


Figure 9-17. Discriminator Circuit

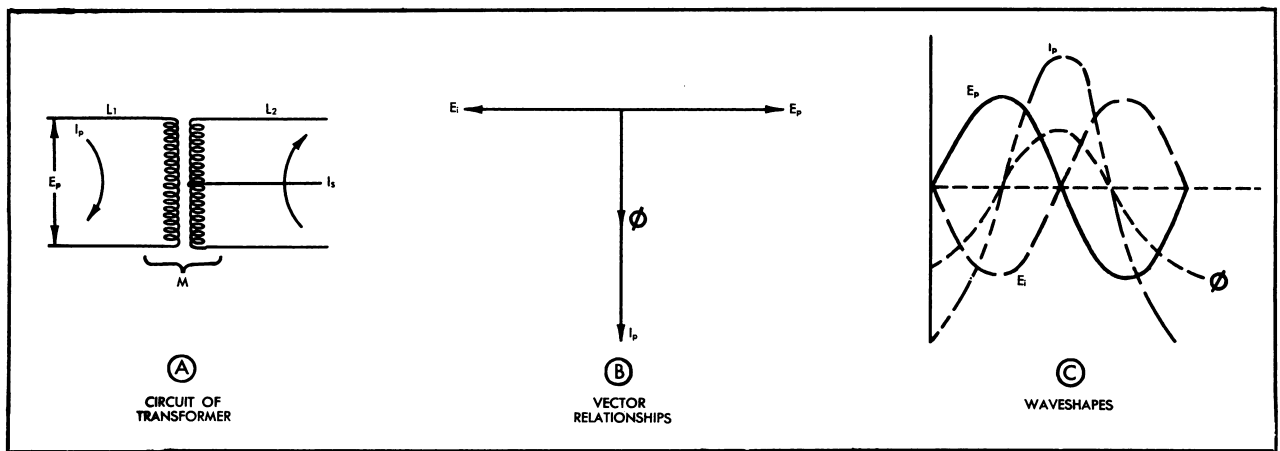


Figure 9-18. Transformer Vector Relationships and Waveshapes

zero output. Therefore, the discriminator is not affected when the amplitude of the applied voltage changes.

The phase relationship of the voltage across the r-f choke to the voltage induced in the secondary is the key to the operation of the discriminator circuit. For this reason you will find it is desirable to analyze its operation in detail. To make the analysis, refer to Figure 9-18 showing the discriminator transformer at A and the vector relationships of the current and voltages in the transformer and their waveshapes.

At the desired frequency the transformer vector relationships are as follows: The primary voltage  $E_p$  is the reference vector. Since the mutual inductance (coupling) is small, the primary is inductive. The primary current  $I_p$  lags  $E_p$  by  $90^\circ$ . The magnetic field which affects the secondary

(called the *flux* and shown by the symbol  $\phi$ ) is in phase with the primary current. Due to normal transformer action, the voltage induced in the secondary is  $90^\circ$  behind the flux.

This induced voltage, labeled  $E_i$ , is simulated by a generator in the equivalent circuit of the secondary shown in Figure 9-19A. The generator is inserted at the center of the transformer secondary where the voltage across the secondary is divided between the two tubes. A current ( $I_s$ ) flows around the loop composed of the secondary windings and capacitor C. This loop is a series-resonant circuit insofar as the generator sees it. Since the circuit is at resonance, the current  $I_s$  is in phase with the induced voltage.

Each half of the secondary has considerable inductive reactance at the radio frequency. Therefore, there is a voltage drop across this

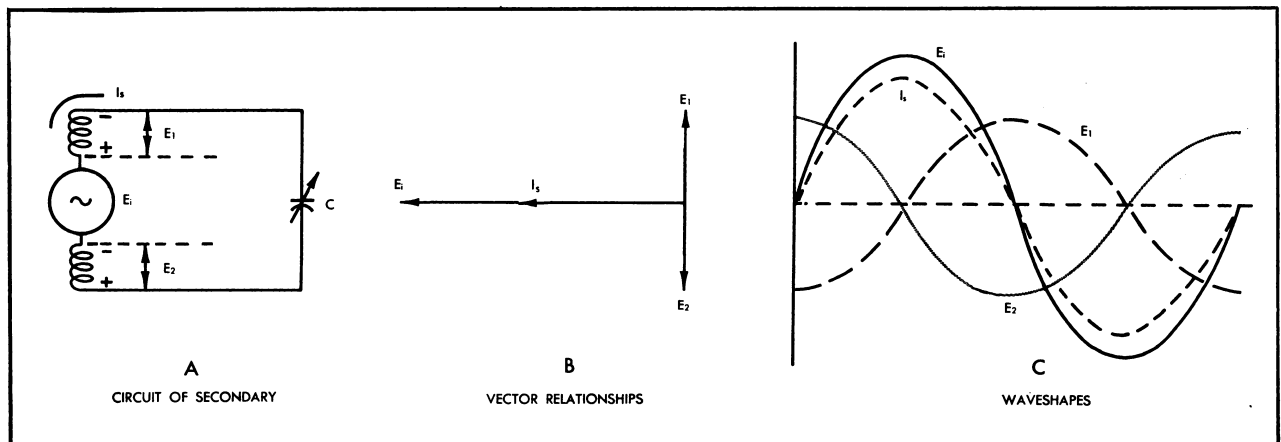


Figure 9-19. Secondary Voltages at Resonance

reactance caused by the current  $I_s$ . These voltages are shown in Figure 9-19A in the secondary circuit as  $E_1$  and  $E_2$ . Since voltage leads current by  $90^\circ$  in an inductor,  $E_1$  and  $E_2$  are  $90^\circ$  out of phase with  $I_s$ . The voltage  $E_2$  actually leads  $I_s$  as it should, but the voltage  $E_1$  is  $180^\circ$  out of phase; that is, its polarity is reversed, due to the manner in which it is connected to the tube. Current flows in the same direction through both coils, but the vector voltages are both measured with respect to the center. This causes opposite polarities to exist.

Note in Figure 9-20A, the equivalent circuit of the discriminator in which the voltages are replaced by generators, that the plate voltage on  $V_1$  is the sum of the voltages  $E_p$  and  $E_1$ , and the plate voltage on  $V_2$  is the sum of  $E_p$  and  $E_2$ . The vector diagram at B shows  $E_p$  and  $E_1$  added vectorially to produce a vector sum  $E_{V1}$ .  $E_{V1}$  is the actual plate voltage on  $V_1$ . The two sine wave voltages  $E_p$  and  $E_1$  are also shown. They add to produce a third sine wave called  $E_{V1}$ . When plate current flows,  $C_1$  charges to near the peak voltage of  $E_{V1}$ , producing a constant DC voltage almost equal to  $E_{V1}$  across resistor  $R_1$ . In the same manner,  $E_p$  and  $E_2$  add, as  $E_p$  and  $E_1$  did previously, resulting in  $E_{V2}$ , the plate voltage on the other diode. Consequently,  $C_2$  charges to  $E_{V2}$  and a constant voltage almost

equal to  $E_{V2}$  appears across  $R_2$  as shown in diagram C. The vector and sine waves are shown at C.

Observe that the voltage across the filters  $R_1C_1$  and  $R_2C_2$  are DC voltages. Obviously, phase relationships cannot exist there. In this case only amplitude relationships are important. Since  $E_p$  is common to both, and  $E_1$  equals  $E_2$ , the vector sums of  $E_{V1}$  and  $E_{V2}$  are equal. The sine-wave illustrations show that the resultants of the sine waves, although of different phase, have equal amplitudes. The DC voltages across each filter are alike, but have opposite polarities. Therefore, the sum of these voltages will be zero between X and Y. This fact indicates that the output at resonance is zero.

#### Output at Frequencies Below Resonance

When the frequency is lower than the center, or resonant, frequency, the output is no longer zero. With the primary voltage as a reference, the voltage induced in the secondary is still  $180^\circ$  out of phase, as shown by the vector diagrams in Figure 9-21A. Since inductive reactance decreases as frequency decreases, while capacitive reactance increases, the secondary circuit of the transformer in Figure 9-19A will be capacitive at frequencies below resonance. The current in a

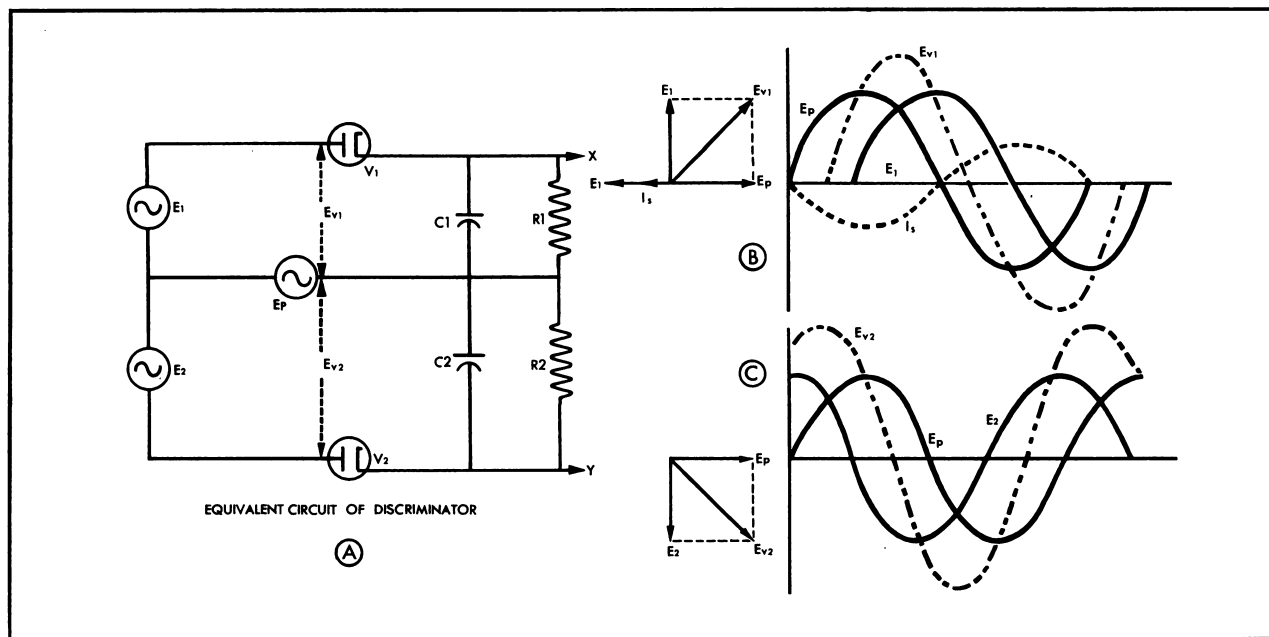


Figure 9-20. Resultant Plate Voltages at Resonance

capacitive circuit leads the applied voltage. Here, the applied voltage is  $E_1$ , and  $I_s$  will lead.

An arbitrary amount of lead is used in the vector diagram in Figure 9-21A. The voltage drops  $E_1$  and  $E_2$  will still be  $90^\circ$  out of phase with the current since this is not a function of frequency. Since under these conditions voltage  $E_1$  is more than  $90^\circ$  away from  $E_p$ , when added vectorially, their sum will be less than at resonance even though the magnitude of the components  $E_p$  and  $E_1$  has not changed. At the same time,  $E_2$  is less than  $90^\circ$  away from  $E_p$  — or they are more nearly in phase — so the magnitude of the resultant  $E_{V2}$  is greater than at resonance. Again, since the resistor voltages are proportional to the length of the resultant vectors, there will be a high voltage across  $R_2$  and a lower voltage across  $R_1$ . Therefore, the output voltage will be equal to the  $R_2$  voltage minus the  $R_1$  voltage and the output point X, Figure 9-20, will be negative with respect to Y which may be grounded.

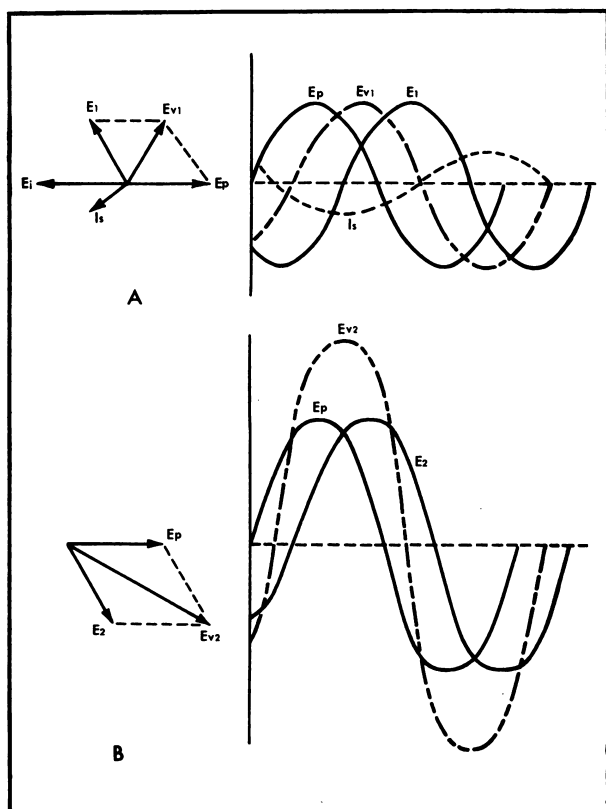


Figure 9-21. Voltages at Frequencies Below Resonance

The waveshape shows the same thing.  $I_s$  is shifted in phase; voltage  $E_1$  is further out of phase with  $E_p$ , and  $E_2$  is more nearly in phase with  $E_p$ . The sum of the more out-of-phase components is small; but the resultant waveshape of the two nearly in-phase components is large.

Studying the vector diagrams, you can see that the greater the deviation from resonance, the greater the DC output voltage.

#### Output at Frequencies Above Resonance

When the input frequency is greater than the center frequency, the circuit  $L_2C$  is inductive, and the current  $I_s$  lags the induced voltage. Figure 9-22 shows the vector relationship of voltages at frequencies above resonance. Note that at frequencies above resonance,  $E_1$  is more nearly in-phase, and  $E_2$  is more out of phase than at the center frequency. Also notice that  $E_{V1}$  has a greater magnitude than  $E_{V2}$ . Therefore, the DC voltage across  $R_1$  is greater than the  $R_2$  voltage, and consequently the output voltage is positive.

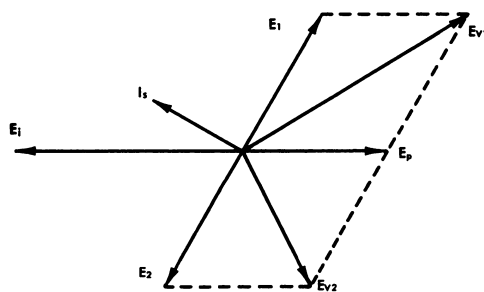


Figure 9-22. Voltages at Frequencies Above Resonance

Remember that a frequency above resonance causes a positive output at X with respect to Y (Figure 9-20). At resonance, the output is zero, and below resonance the output is negative. These polarities may be reversed by measuring the voltage at Y with respect to X (Figure 9-20).

#### Use

The discriminator circuit is used as a part of the automatic tuning circuit in a receiver. Part of the i-f signal is diverted to the input transformer of the discriminator. The output of the discriminator is connected to the local oscillator of the receiver in such a way that the DC volt-



age affects the frequency of the oscillator. If the intermediate frequency is correct, the receiver is operating at peak efficiency, the discriminator has a zero output, and the oscillator frequency is not affected. If the i-f changes, the receiver is said to be detuned, and the i-f section operates at lower gain. However, the discriminator develops a DC voltage from this incorrect frequency and this DC voltage changes the frequency of the oscillator. By proper circuit design, this change can be made to occur in the direction that will correct the intermediate frequency.

The oscillator, of course, beats with the incoming r-f signal to form the i-f signal. Therefore, a change of oscillator frequency, or a change of input frequency, changes the intermediate frequency. The discriminator and its associated circuit corrects either change. This system is known as *automatic frequency control*, and is usually abbreviated as AFC.

### AUTOMATIC FREQUENCY CONTROL

The output of a discriminator is normally applied to a loudspeaker or to some indicating device. However, a discriminator used in conjunction with a reactance tube and oscillator may be used in a circuit to achieve automatic frequency

control. Automatic frequency control used in AM or FM receivers is usually referred to as AFC, but in radar circuitry the same device may be referred to as a *phase detector* or *resolver*. Circuit operation in either case is identical but their applications may vary.

In the block diagram of Figure 9-23, the sine-wave oscillator, which is operating at or near the desired constant output frequency, delivers an output in two directions: one is called the constant frequency output; the other, a sample signal, is applied to the discriminator where it is compared with the incoming carrier frequency. Slight variations in frequency or phase between the carrier frequency and the sine-wave output of the oscillator produce a voltage output from the discriminator that indicates the magnitude of phase difference between the two signals.

The balanced diode discriminator circuit produces a change in rectified output when the frequency or phase of the carrier frequency changes. Therefore, any difference in frequency between the carrier frequency and the sine-wave oscillator frequency is measured by the discriminator. The discriminator output can then be used to adjust the sine-wave oscillator frequency with the reactance tube, bringing the two signals back into phase and frequency to produce the desired constant frequency output (with respect to the input carrier).

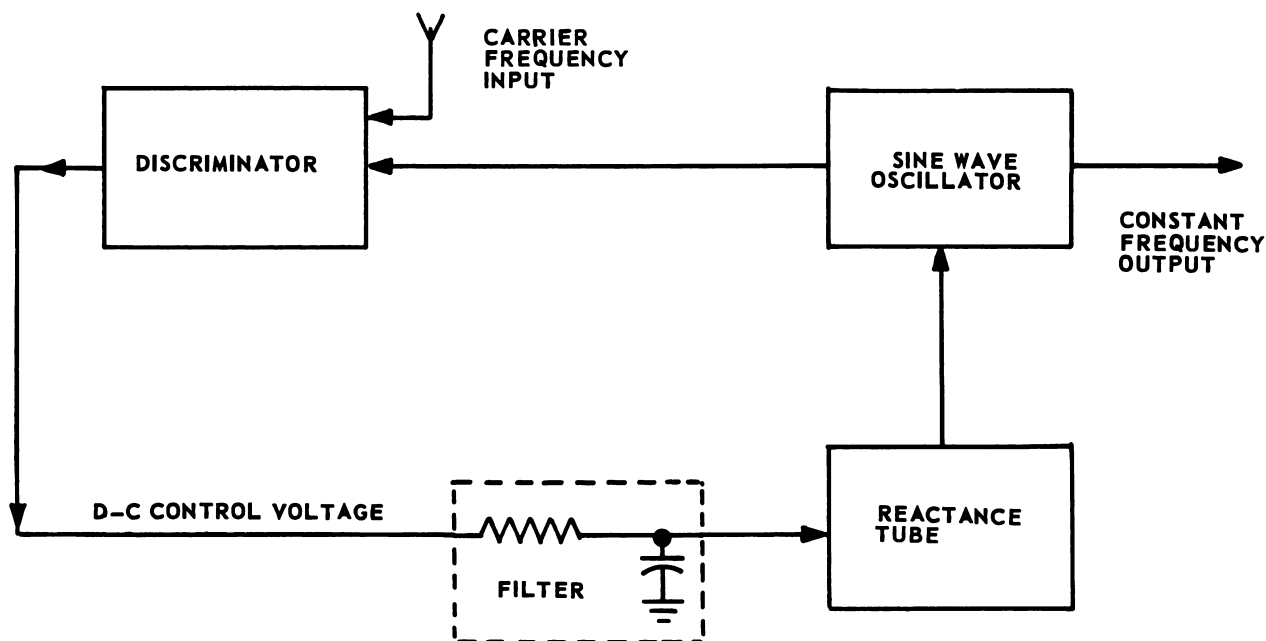


Figure 9-23. Automatic Frequency Control Circuit Block Diagram

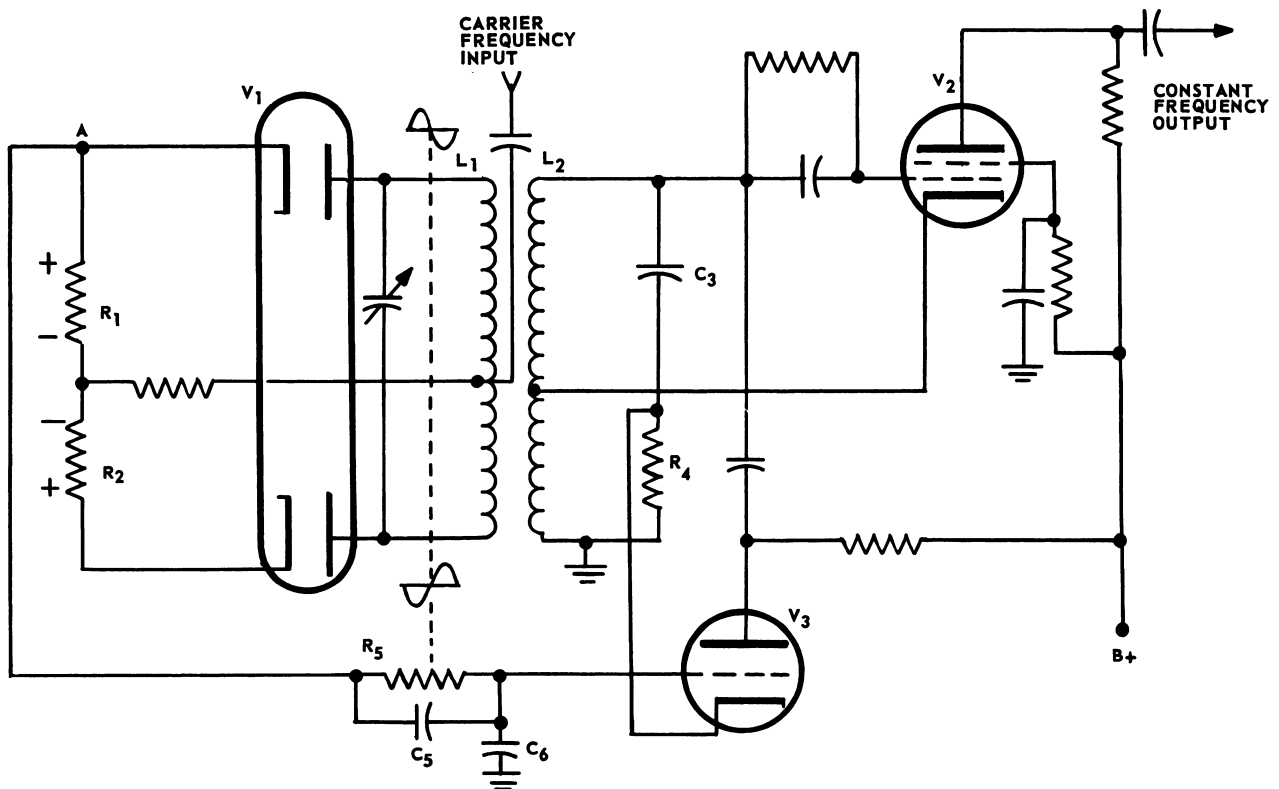


Figure 9-24. Schematic Diagram of an Automatic Frequency Controlled Circuit

Figure 9-24 shows an electron-coupled Hartley oscillator circuit ( $V_2$ ), with the screen grid acting as the oscillator anode. The oscillator coil  $L_2$  is closely coupled to the center-tapped discriminator coil  $L_1$ , applying equal voltages in opposite phase to the diode plates of the discriminator. This creates a zero difference voltage (DC control voltage) across  $R_1$  and  $R_2$  which is applied to the grid of the reactance tube ( $V_3$ ).  $C_3$  and  $R_4$  provide the required quadrature network, with the voltage across resistor  $R_4$  coupled to the cathode circuit instead of the control grid so that the DC control voltage from the discriminator can be coupled to the control grid of  $V_3$ . Rapid changes in the DC level of the discriminator output that might be produced by interfering noise signals are attenuated by the RC filter network ( $R_5$ ,  $C_5$ , and  $C_6$ ).

If the carrier frequency is above its correct value (the resonant frequency of the discriminator circuit), the resulting DC control voltage between point A and ground is positive. This positive control voltage is applied to the control grid of the reactance tube ( $V_3$ ) and causes the oscillator frequency to increase to its proper

value. When the oscillator is again operating at the proper frequency, the discriminator output voltage is zero, and therefore the reactance tube causes no further change in frequency. If the carrier frequency is below its correct value, the resulting DC control voltage between point A and ground is negative. This negative control voltage, which is applied to the grid of the reactance tube, causes the oscillator frequency to decrease to its proper value. A similar action occurs when the sine-wave oscillator tends to fall above or below the carrier frequency.

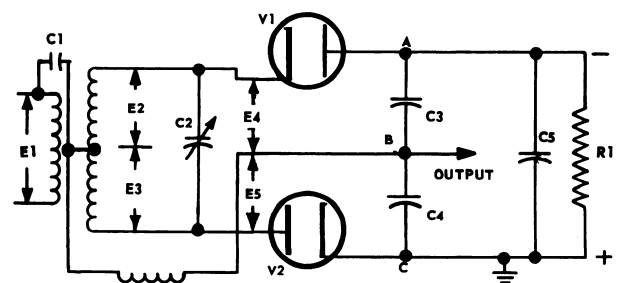


Figure 9-25. Ratio Detector

## RATIO DETECTOR

The output of nearly all frequency modulated oscillators also varies in amplitude to some degree. The input to the discriminators discussed so far should be of constant amplitude, varying only in frequency. Otherwise, the output of the discriminator will be distorted; it will not vary exactly in accordance with the modulating signal that produced the frequency variations. A limiter circuit must be used to eliminate the amplitude modulation before the signal is applied to these discriminators.

The ratio detector, shown in Figure 9-25, is a circuit designed to respond only to frequency changes of the input signal. Amplitude changes in the input have no effect upon the output. The input circuit of the ratio detector is identical to that of the discriminator circuits previously covered in this chapter. It uses a center-tapped transformer secondary as they did.

The voltages appearing across  $V_1$  ( $E_4$ ) and  $V_2$  ( $E_5$ ) when the input is below, at, and above the center frequency are shown in Figure 9-26. The input circuit is the only portion of the ratio detector that operates in the same way as the discriminators already discussed.  $V_1$ ,  $R_1$ , and  $V_2$  form a series circuit that is fed by the tuned circuit of the transformer secondary. Since the two diodes are connected in series, they conduct on the same half-cycle of the input r-f.

The rectified current flowing through  $V_1$  and  $V_2$  produces a voltage drop across  $R_1$  with a nega-

tive polarity at the plate of  $V_1$  as shown in Figure 9-25.  $C_3$  is a very large capacitor connected in parallel with  $R_1$ . Its function is to keep the voltage drop across  $R_1$  constant even for the slowest frequency deviations of the input signal. Thus, the voltage between points A and C, Figure 9-25, remains constant. This is made possible by the large value of  $R_1$  and  $C_3$ . This is the reason amplitude variations in the input voltage have no effect on the output.

The rectified voltage across  $C_3$  is proportional to  $E_4$ , the voltage appearing across  $V_1$ . The rectified voltage across  $C_4$  is proportional to  $E_5$ , the voltage appearing across  $V_2$ . It has been pointed out that  $E_4$  and  $E_5$  vary in magnitude in accordance with the frequency variations of the input signal. Thus, the voltage across  $C_3$  and that across  $C_4$  vary in the same manner. The sum of the voltages across the two (between points A and C) remains constant. However, the output taken at point B varies with respect to points A and C. If the potential across  $C_3$  increases, that across  $C_4$  decreases by the same amount, since their sum is fixed by the constant voltage across  $C_5$ . This circuit gets the name ratio detector because the ratio of the voltage across  $C_3$  to that across  $C_4$  may be continually varying, while the sum of the two voltages remains constant.

With the circuit connected as shown in Figure 9-25 the output is always taken with respect to ground and varies about an average negative potential. The output voltage, as the frequency

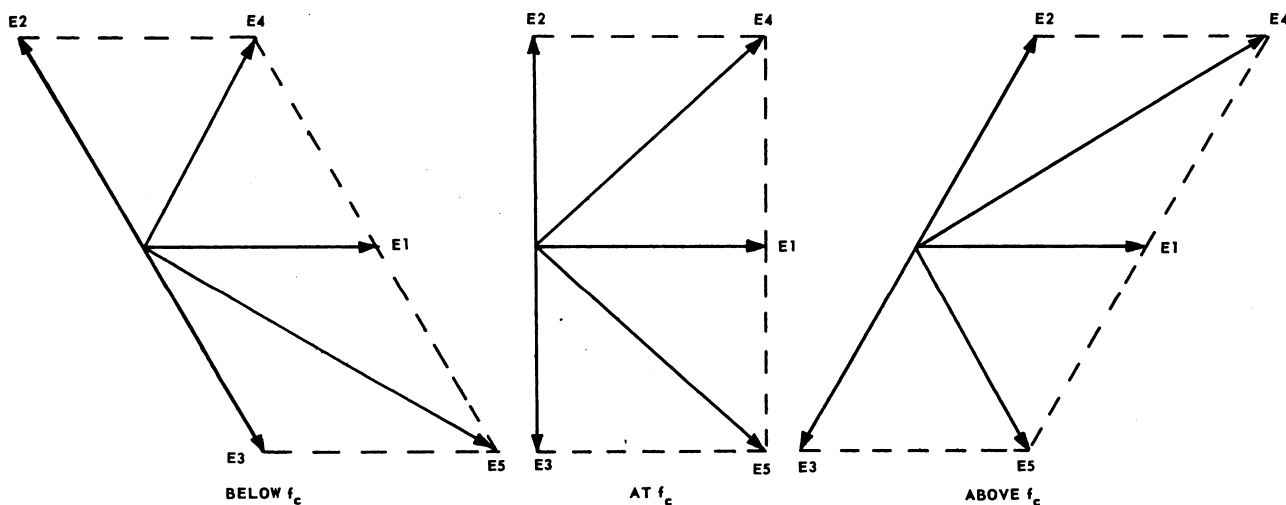


Figure 9-26. Voltage Relationships at Input of Ratio Detector

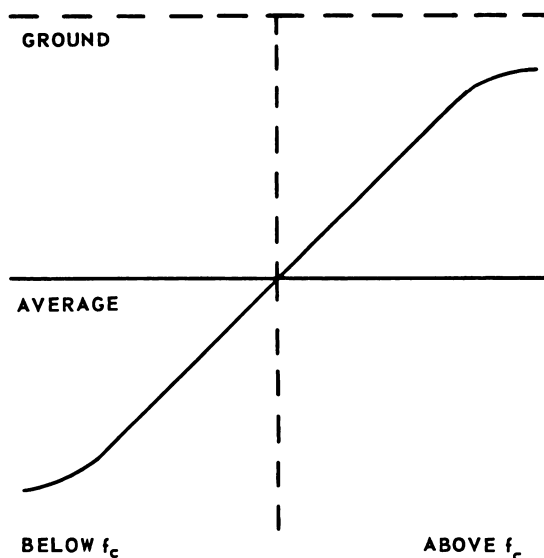


Figure 9-27. Output of Ratio Detector

varies, can be determined from Figure 9-26. Remember, that the voltage across  $C_3$  is proportional to  $E_4$  and that across  $C_4$  is proportional to  $E_5$ . When the input frequency is below  $f_c$ ,  $E_5$  is greater than  $E_4$ , producing a voltage across  $C_4$  that is greater than that across  $C_3$ . At this time, then, the output is maximum negative with respect to ground, as shown in Figure 9-27. When the input is at  $f_c$ , the voltages across  $C_3$  and  $C_4$  are equal, since  $E_4$  equals  $E_5$ . At this time the output is at the average negative potential. When the input frequency is above  $f_c$ , the voltage across  $C_3$  is greater than that across  $C_4$ , since  $E_4$  exceeds  $E_5$ . The output at this time is the least negative, as shown in Figure 9-27.

The ratio detector is becoming more and more popular, since its use eliminates the necessity of a limiter stage in FM receivers.

### SINGLE-SIDEBAND TRANSMISSION

The basic concept of single-sideband (SSB) communications was understood as early as 1914. It was first realized through mathematical analysis of an amplitude-modulated r-f carrier. It can be shown that a carrier modulated by a single sine wave of voltage consists of three different frequencies. These are: (1) the original carrier with amplitude unchanged; (2) a frequency equal to the difference between the carrier and the modulating frequencies, with an amplitude

equal to one-half that of the modulating signal; and (3) a frequency equal to the sum of the carrier and the modulating frequencies, with an amplitude equal to one-half that of the modulating signal. The two new frequencies are called *sidebands*.

Upon recognition of the fact that sidebands existed, further studies were made. These investigations and experiments proved that after the carrier and one of the sidebands were eliminated, the other sideband could be used to transmit the intelligence. (Since the carrier amplitude is unchanged, there is no intelligence contained in the carrier.) Further experiments proved that both sidebands could be transmitted, each containing different intelligence, with a suppressed or completely eliminated carrier.

By 1923 the first patent for this system had been granted, and a successful SSB communications system was established between the United States and England. Today SSB communications play a vital role in providing dependable and efficient communications between the United States and its Armed Forces scattered throughout the world.

To gain a thorough understanding of single sideband communications, you must first understand the principles of amplitude modulation. When two different signals are applied to a non-linear device, heterodyning action occurs. The output of this device will contain the upper sideband (USB), the lower sideband (LSB) and the carrier frequencies.

Let us look at the composite AM waveform from a frequency spectrum standpoint. In Figure 9-28, one represents the carriers, two the LSB, and three the USB. Assume that the constant amplitude 100-kc r-f carrier is modulated by a voice frequency signal that is varying in frequency from 100 to 6000 cycles and is changing in amplitude. The output of this device will contain difference frequencies ranging from 94 to 99.9 kc (LSB) and sum frequencies ranging from 100.1 to 106 kc (USB). Notice that each sideband has the same frequency separation from the carrier. They contain the same intelligence, which can be extracted from either sideband at the receiver.

The sidebands vary in a horizontal direction (left-right, Figure 9-28B) from the carrier. As the modulating frequency varies, they move closer to, or farther away from the carrier, the

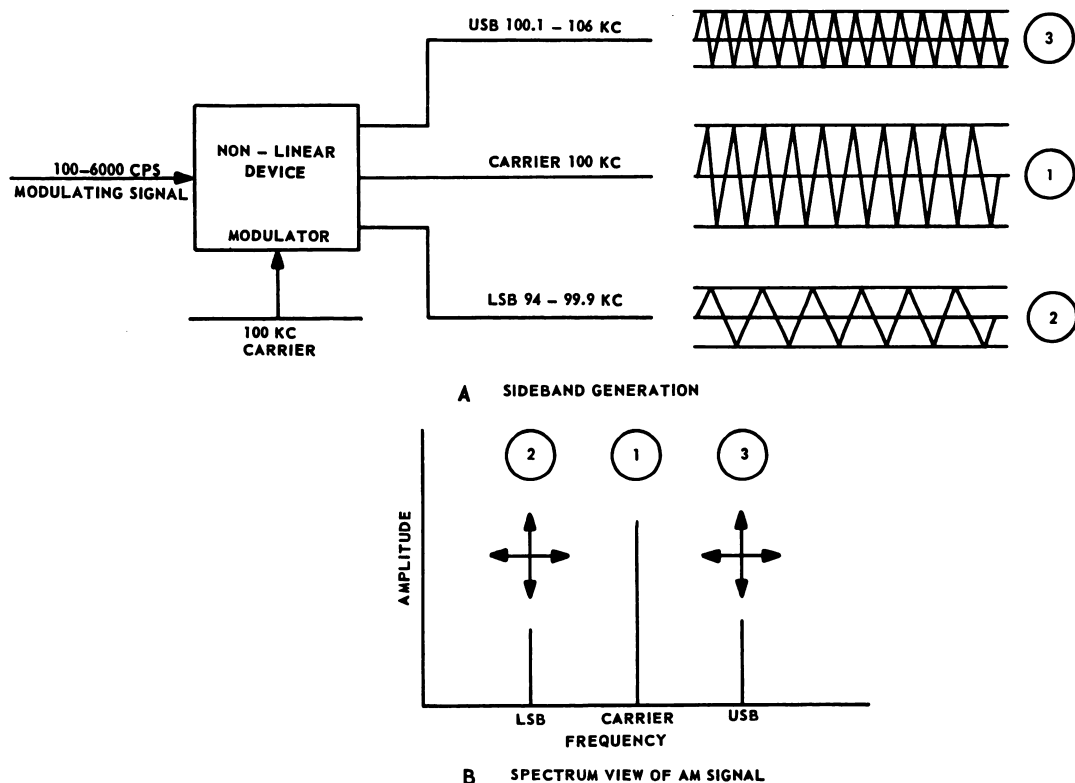


Figure 9-28. Amplitude Modulation

amount depending upon the low and high frequencies of the modulating signal. The amplitude of the sidebands varies, depending upon the volume level of the modulating signal. In voice modulation, speaking in a loud tone increases the amplitude of the sidebands; soft tones decrease the amplitude. From this you can see that the sidebands are continuously moving left and right (frequency) and up and down (amplitude), as shown in Figure 9-28B.

These sideband frequencies are generated as a result of the mixing of the r-f carrier and the modulating signal in a nonlinear device. The radiated signal consists of three frequency components (carrier, USB and LSB) of a definite amplitude and frequency relationship.

When we view the radiated signal on an oscilloscope, we see a pattern that is referred to as a composite AM waveform. The outline of this pattern is called the *modulation envelope*. It will always resemble the original modulating signal, if no distortion is introduced. A 100-kc carrier modulated by a 6-kc tone will appear as shown in Figure 9-29. The composite AM waveform appears to be a radio-frequency carrier

varying in amplitude. It is the pattern of the instantaneous frequency-voltage relationship of the carrier and the sideband frequencies, or the vector sum of their three frequency components. It can also be said to be the over-all resultant signal voltage of the phase relationship of the three frequency components at one given instant. The carrier, upper sideband, and lower sideband power and frequency are distributed throughout this waveform. That is, there is as much carrier, upper, and lower sideband power and frequency at point X as there is at points Y and Z.

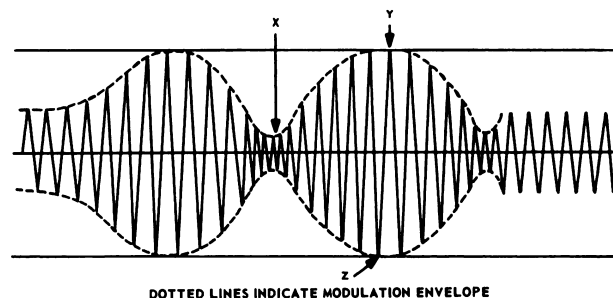


Figure 9-29. Composite AM Waveform

## Percentage of Modulation

The amplitude of the sidebands generated is always proportional to the amplitude of the audio or modulating signal. The amplitude of this modulating signal will also determine the percentage of modulation. This term is used to describe the power distribution of the carrier and the sidebands; it will determine the usability of the transmitted signal. One-hundred percent modulation takes place when the system is delivering maximum undistorted power output. With less than 100% modulation, the intelligence in the envelope is weaker. With more than 100% modulation, excessive distortion results. The percentage of modulation, therefore, is quite important. Just as sideband generation depends upon the audio or modulating signal, so does the percentage of modulation.

When two equal voltages of different frequencies are applied to a nonlinear device, the voltage at the output, or plate, of the stage will vary

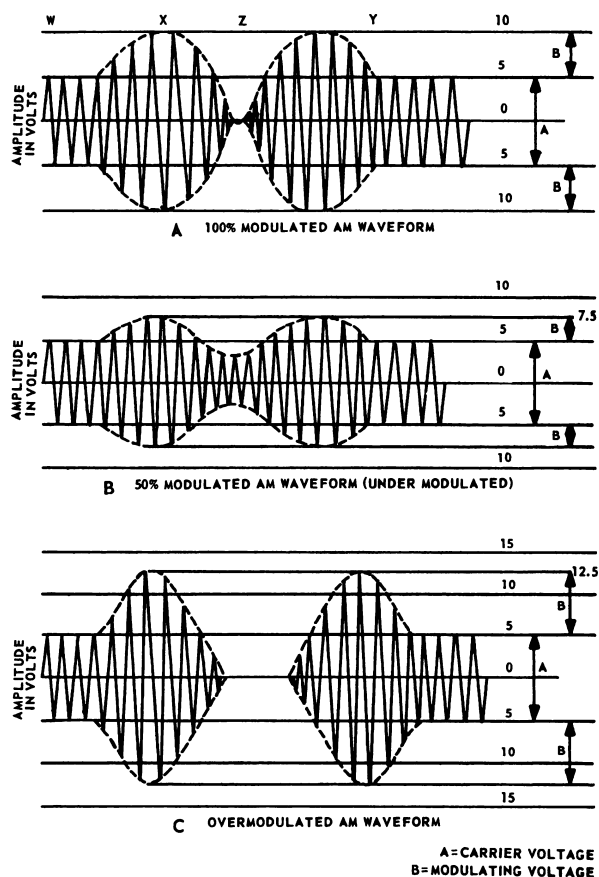


Figure 9-30. Percentage of Modulation

from twice its normal value to zero, depending upon whether the two voltages are in phase or out of phase. Because of the frequency difference between the two signals, it will be impossible for them to be completely in or out of phase at all times. This explains the oscilloscope presentation of the composite AM waveform, where the carrier *appears* to be varying at an audio rate. The sidebands that are generated form a definite phase relationship with respect to the carrier. Once this relationship is formed, it must be maintained all the way to the detector in the receiver. Otherwise, severe distortion will occur, and in some cases, complete loss of carrier or signal intelligence.

Figure 9-30 shows examples of modulated AM signals. Figure 9-30A represents a 100% modulated wave. Notice that the section labeled W is of constant amplitude. This is the carrier before it has been modulated. Since the carrier and modulating voltages are of equal amplitude, the carrier appears to increase in amplitude until at point X the amplitude of the composite waveshape is twice the unmodulated carrier. At Z the amplitude has decreased to zero. After one and one-half cycles of the modulating signal, the carrier is again unmodulated and of constant amplitude.

In Figure 9-30B, the modulating voltage amplitude is less than that of the carrier. The composite waveshape does not increase as much as in the first example, nor does the amplitude ever drop to zero. Figure 9-30C shows a carrier modulated by a signal of greater amplitude. The peak voltage is greater than twice the unmodulated carrier, while the output drops to zero at an earlier point than in example A and remains there for a period of time. The envelope in Figure 9-30C does not resemble the original modulating signal.

The percentage of modulation can be figured by using the following formula.

$$\text{Percentage of modulation} = \frac{E_{\max} - E_{\min}}{E_{\max} + E_{\min}} \times 100 \quad (1)$$

where  $E_{\max}$  is the maximum voltage of the modulated wave and

where  $E_{\min}$  is the minimum voltage of the modulated wave.

Another way of finding the percentage of modulation is,

$$\text{Percentage of Modulation} = \frac{\text{Peak amplitude of modulating voltage}}{\text{Peak amplitude of carrier voltage}} \times 100 \quad (2)$$



### Examples

Find the percentage of modulation for the waveshapes in Figure 9-30, using both formulas where applicable.

For those in Figure 9-30A:

$$\frac{E_{\max} - E_{\min}}{E_{\max} + E_{\min}} \times 100 \quad (1)$$

$$\frac{10 - 0}{10 + 0} \times 100 = 100\%$$

or  $\frac{\text{Peak amplitude of modulating voltage}}{\text{Peak amplitude of carrier voltage}} \times 100 \quad (2)$

$$\frac{5}{5} \times 100 = 100\%$$

For those in Figure 9-30B:

Using formula (1),

$$\frac{7.5 - 2.5}{7.5 + 2.5} \times 100 =$$

$$\frac{5}{10} \times 100 = 50\%$$

Using formula (2),

$$\frac{2.5}{5.0} \times 100 = 50\%$$

For those in Figure 9-30C:

Formula (1) cannot be used for overmodulation (greater than 100% modulation).

Using formula (2),

$$\frac{7.5}{5.0} \times 100 = 150\%$$

### Power Distribution

In the composite AM waveform, there are two values of power to be considered: the carrier power, and the peak power. Carrier power is the average power in the unmodulated carrier. This is point W in Figure 9-30. Peak power is the power developed when the carrier and both sidebands are in phase. If we know the voltages, we can calculate the power in any of the components of the AM waveform.

If we have a carrier modulated 100% by a single sine wave of voltage of constant amplitude, the computations are not too difficult. Assume that the rms voltage of the unmodulated carrier is 100 volts and the radiation resistance is 50 ohms. The power in the unmodulated carrier can be found from the power formula,

$$P = \frac{E^2}{R}$$

$$P = \frac{(100)^2}{50} = \frac{10000}{50} = 200 \text{ watts.}$$

Since the amplitude of the carrier remains constant, modulated or unmodulated, the power in the carrier is 200 watts at all times.

For 100% modulation, the modulating signal voltage is equal to the carrier voltage, in this case 100 volts. Then the rms value of the sideband voltage is 50 volts. The power in either sideband can be found from the power formula,

$$P = \frac{(50)^2}{50} = 50 \text{ watts.}$$

As you can see, the power in each sideband is one-fourth the power in the carrier at 100% modulation. The total sideband power is 100 watts. With a carrier power of 200 watts, this gives a total power of 300 watts. This means that to attain 100% modulation of an r-f carrier with a sine wave modulating signal, a modulating power equal to one-half the carrier power is required. The additional modulating power is divided equally between the upper and lower sidebands.

The peak power for this 100% modulated wave can be found as follows. The rms value of  $E_{\max}$  is 200 volts. Using the power formula,

$$P = \frac{(200)^2}{50} = \frac{40000}{50} = 800 \text{ watts.}$$

While all of the intelligence is contained in the sidebands, two-thirds of the total power is in the carrier. It would appear that a great amount of power is wasted during transmission. The basic principle of single sideband transmission is to eliminate or greatly suppress the high energy r-f carrier. This can be accomplished without affecting the fidelity of the emitted intelligence, since the carrier contains no intelligence.

If a means of suppressing or completely eliminating the carrier is devised, the power that was used for the carrier can be converted into useful power to transmit the intelligence in the sidebands. Since both upper and lower sidebands contain the same intelligence, one of these could also be eliminated, thereby adding more power for transmission of the other intelligence-bearing sideband. This would result in more power in the radiated signal and a greatly extended range of transmission.

### Power Comparison

The total power output of a conventional AM transmitter is equal to the carrier power plus

the power output of the modulator. Conventional AM transmitters are rated in carrier power output. In single sideband operation, there is either a greatly attenuated carrier or a completely eliminated carrier. For this reason, the SSB transmitter is rated in peak envelope power. In the example of the 200-watt transmitter previously used, the peak power was 800 watts. If this same transmitter were converted to SSB operation, it would be rated as an 800-watt single sideband transmitter, and a 6-db gain in power would be realized. (Any power doubled equals a 3-db gain.) This gain is due to the elimination of the carrier and one of the sidebands.

The bandwidth of the 800-watt SSB transmitter would be only one-half that required for a conventional AM transmitter. This is an advantage that results in 3-db gain at the receiver. This gain is due to the better signal-to-noise ratio obtained. Another advantage realized is that the sideband signal is less subject to the effects of selective fading, which is caused by the out-of-phase relationship between the carrier and sidebands. It commonly occurs in conventional AM transmission. In this type of single sideband system, a theoretical 9-db gain is realized over the conventional AM system.

### Types of Sideband Transmission

Several types of single sideband systems have been developed. Therefore, it is necessary to clarify the various applications of the transmission principle involved. The term *single sideband modulation* is defined as a form of amplitude modulation in which one sideband and the carrier are suppressed.

In one system, the carrier and one of the sidebands are eliminated at the transmitter; only one sideband is emitted. This is called *single sideband suppressed carrier* (SSSC); it is quite popular with amateur radio operators. The chief advantage of this system is maximum transmitted signal range with minimum transmitter power.

Another system eliminates one sideband and suppresses the carrier to a desired level. The suppressed carrier is then used at the receiver for a reference, AVC, AFC, and in some cases, demodulation of the intelligence-bearing sideband.

The type of system most commonly used in Air Force communications is referred to as *twin sideband suppressed carrier*, or *independent sideband* transmission. This system involves the

transmission of two independent sidebands, each containing different intelligence, with the carrier suppressed to a desired level.

### Advantages of SSB Transmission

The most important advantage of the independent sideband system is a more effective utilization of the available frequency spectrum. In the same bandwidth required for the transmission of one conventional AM signal, with present day multiplexing techniques, many channels of different intelligence can be simultaneously transmitted by one SSB transmitter within one assigned operating frequency. This makes this type of communications especially adaptable to the already overcrowded high-frequency spectrum.

A second advantage of this system is that it is less subject to the effects of selective fading. This is because there is no definite phase relationship between the upper and lower sidebands and the carrier as there is in conventional AM. In the propagation of conventional AM transmissions, if the upper sideband frequency strikes the ionosphere and is refracted back to earth at a different angle from that of the carrier and lower sideband frequencies, distortion is introduced at the receiver. Under extremely bad conditions, complete signal cancellation may result, which means a complete loss of the intelligence.

Any change in the phase relationship between the sidebands and the carrier during the transmission of the conventional AM signal will result in distortion at the receiver during demodulation. If the upper sideband frequency of an independent sideband transmission is refracted back to the earth at a different angle from that of the carrier and lower sideband, it will cause the signal strength for that sideband to decrease. There will be no effect on the other sideband or the suppressed carrier whatsoever; thus, only a partial loss (in the USB) of the entire intelligence transmitted occurs.

Another advantage realized by all types of SSB systems is that a higher percentage of power is in the radiated intelligence. The independent sideband system has an estimated over-all 6-db gain, while the single sideband suppressed carrier system (SSSC) has an over-all 9-db gain, 6 db at the transmitter and 3 db at the receiver. In

the example in the text, the 200-watt conventional AM transmitter was converted to a SSSC SSB transmitter, and an over-all 9-db gain was realized. If this same transmitter were used for independent sideband operation, we would realize only an over-all gain of 6 db (3 db at the transmitter and 3 db at the receiver). The power would be equally divided between the upper and lower sidebands. That is, there would be approximately 400 watts in each sideband. This, then, represents twice the value of our original 200-watt reference, or a 3-db increase at the transmitter.

### Receiver Considerations

When considering the gain of any system, the receiver must be taken into consideration. In the conventional AM system, the signal applied to the detector consists of the upper and lower sidebands with a full-power carrier. However, due to the rectification by the diode detector, only the positive portion of the signal is used to develop the output. Therefore, we have wasted half the available power in the signal.

In most SSB systems used by the Air Force, each full sideband is detected or demodulated by its own full-wave demodulator. Since all of the power in each sideband signal is applied to a different demodulator, we realize a 3-db gain at the receiver. Another advantage contributing to the 3-db gain of the receiver is the narrow bandwidth. Since the bandwidth required for each sideband is one-half that required for conventional AM, a higher signal-to-noise ratio is possible. As propagation conditions become worse because of fading or other atmospheric interferences, the signal-to-noise ratio improvement of the SSB system over an equivalent AM system becomes increasingly greater.

### Disadvantages of SSB

One of the disadvantages of SSB equipment is its high cost. This results from the complex circuitry and type of components used. Special emphasis is placed upon voltage regulation, stability, and reliability. The oscillator frequencies involved are very critical and must be stabilized; otherwise, distortion will occur. In SSB systems every precaution is taken to prevent frequency drift, because when the carrier drifts, the sidebands will drift proportionately. If corrections

are not made, the intelligence will be distorted upon demodulation.

The suppressed carrier is amplified and filtered for use in special circuits such as AVC and AFC at the receiver. In some cases this *reconditioned carrier* is also used to demodulate the sidebands of intelligence. By using these special circuits, any frequency drift of the carrier will be corrected by the AFC; variations of the input signal level will be compensated for by the AVC, thereby maintaining a constant output to overcome the effects of fading. The reconditioned carrier may be used to demodulate the sidebands; however, due to propagation conditions, this signal is usually varying in both frequency and amplitude, and therefore is unsuitable for this purpose.

In some systems the reconditioned carrier must be used in the demodulation process; however, in normal communications systems, a highly stable local oscillator is used for demodulation. This highly stable reference oscillator is used at the receiver to generate the conversion frequencies, which will provide distortionless demodulation of the signal intelligence.

The importance of AVC, AFC and the reconditioned and local carriers cannot be overemphasized. When demodulating the sideband signal in the receiver, the final conversion frequency must be within 15 cycles of the transmitter carrier for good quality voice reproduction. It must be within 5 cycles for peak performance of frequency-shift tone teletype signals, and within 1 cycle for the highly sophisticated multitone systems.

### Sideband Generation

The techniques of generating sidebands will vary with different equipment designs. In the conventional AM systems, the sidebands are generated at the final PA stage. In an SSB system, they are generated preceding the final stage. Also in the AM system the fundamental or carrier frequency is produced by an oscillator and then applied to a series of multiplier stages until the final operating frequency is obtained. At this point, the r-f carrier is modulated by the audio-frequency intelligence and applied to the antenna.

The basic principle of SSB requires the suppression or elimination of the r-f carrier and

one of the sidebands. Therefore, specially designed circuits are used to perform these functions at a low-power level. Modulation at the low level necessitates the use of linear power amplifiers rather than frequency multipliers in order to keep distortion to a minimum. The most common method used to generate sidebands is frequency conversion coupled with selective filtering.

Since the primary objective is to convert the audio intelligence into a sideband frequency, it is necessary that a means be devised to eliminate the carrier and one of the sidebands. This is easily accomplished by using a *balanced modulator circuit*. The main characteristic of a balanced modulator is that the carrier or conversion frequency is canceled in the output of the modulator. Therefore, the output contains only the upper and lower sidebands of intelligence. The undesired sideband may be eliminated by using a selective filter, leaving only the desired sideband for transmission. This signal is applied to linear amplifiers and further converted until the final operating frequency is obtained.

### Balanced Modulators

Before single sideband, suppressed-carrier transmission can occur, two things must be accomplished. First, the carrier must be eliminated, or suppressed; and second, one of the two sidebands must be filtered out. Once this has been

achieved, the selected sideband can be applied to the final power amplifier for transmission. A *balanced modulator* circuit can be used to pass only the two sidebands. Balanced modulation is a system of adding intelligence to a carrier wave whereby only the sidebands are produced; the carrier wave is eliminated.

The balanced modulator resembles the conventional push-pull amplifier in circuitry but not in operation. Figure 9-31 shows a conventional push-pull amplifier. For proper operation as an amplifier the signals applied to the grids of  $V_1$  and  $V_2$  must be  $180^\circ$  out of phase. When the positive cycle is applied to the grid of  $V_1$ , plate current flowing through  $T_2$  will produce the positive portion of the output voltage. When the grid of  $V_1$  goes negative, the grid of  $V_2$  goes positive and plate current from  $V_2$  flows in the opposite direction through the primary of  $T_2$ , producing the negative-going alternation of the output waveshape.

By slightly modifying the push-pull amplifier, as shown in Figure 9-32, no output signal will be obtained when an input signal is applied. Notice that the input signal applied to the grids of  $V_1$  and  $V_2$  in Figure 9-32 has the same polarity. Both grids are positive at the same time. A positive-going grid in each tube causes the plate current to increase simultaneously in each tube. Since the plate current of  $V_1$  is flowing down through the primary of  $T_2$  at the time the plate

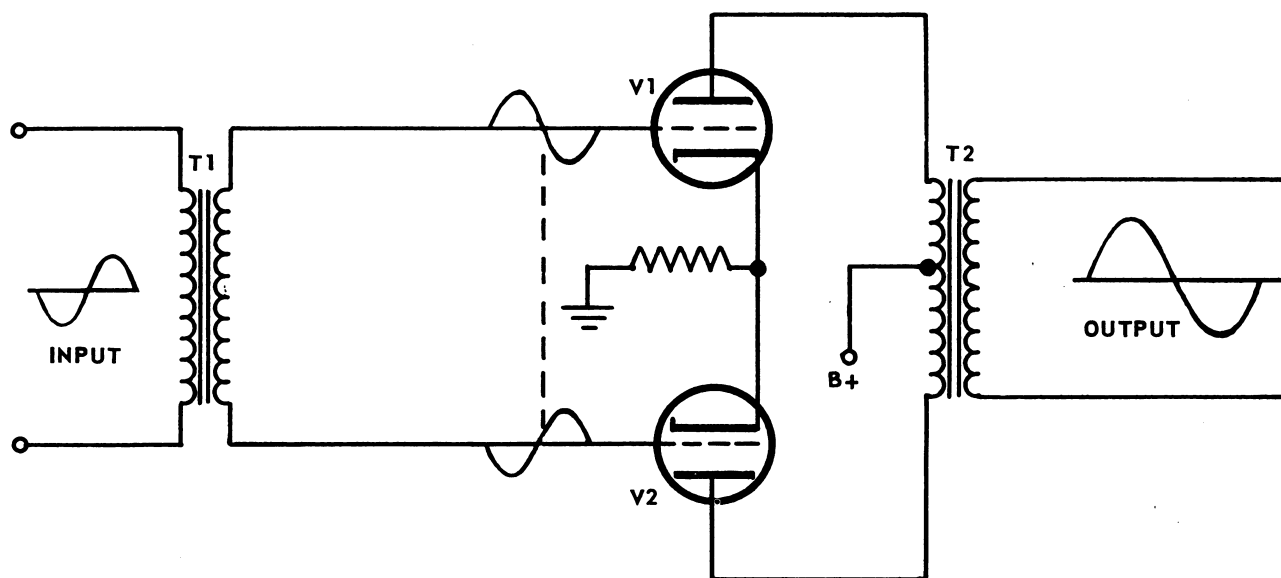


Figure 9-31. Conventional Push-Pull Amplifier

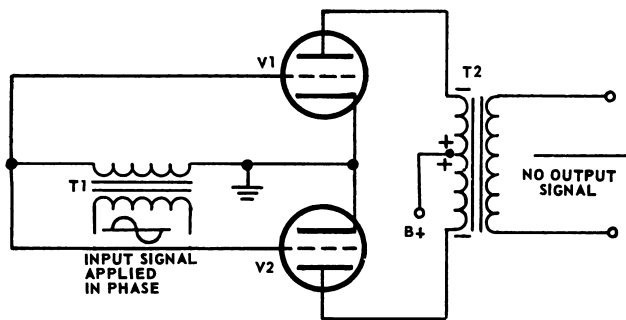


**Figure 9-32. Push-Pull Amplifier with Signal Applied in Phase**

current of  $V_2$  is flowing up, the two magnetic fields produced around  $T_2$  will effectively cancel. Thus, no output signal will be developed. At first glance, this particular circuit is not performing any useful function. On the other hand, assume the input signal to be the carrier frequency. When the carrier frequency is applied in phase, no output signal will be developed in the output; thus, the carrier frequency has been suppressed, or eliminated, in the output.

Again modify the push-pull amplifier circuit and get the circuit shown in Figure 9-33 (essentially a combination of Figures 9-31 and 9-32). In this simplified push-pull balanced modulator, the carrier input is applied to the grids of both  $V_1$  and  $V_2$  in phase; when the r-f voltage is going positive, both grids are driven positive. The modulating voltage is applied to the two grids in the conventional push-pull manner; when the

modulating signal drives the grid of  $V_1$  in a positive direction, the grid of  $V_2$  is driven in a negative direction.

Now consider the effect of the r-f voltage on the stage. Since the carrier frequency causes both grids to be driven positive at the same time, the plate current of both  $V_1$  and  $V_2$  increases at the same time. Consequently, the magnetic fields caused by the two equal plate currents are effectively cancelled in the output transformer  $T_2$ , which is center tapped to  $B+$ . On the negative swing of the r-f voltage, the plate current of both tubes decreases at the same time. Again, the magnetic fields cancel. The carrier frequency, therefore, does not appear in the output.

The effect of the modulating voltage is different. When the positive swing of the modulating voltage is applied, the grid of  $V_2$  is driven negative. On the negative alternation of the modulating voltage, the grid of  $V_1$  is driven negative, but the grid of  $V_2$  is driven positive. Because of this push-pull arrangement, plate current flows on both the positive and the negative swings, first through one tube and then through the other. There is no cancellation in the output.

When the modulating and r-f voltages are applied at the same time, first one tube conducts the most and then the other, depending upon which grid is driven positive by the modulating voltage. The modulating voltage acts as a varying bias. When the grid of  $V_1$  is positive, r-f currents at the carrier frequency, the modulat-

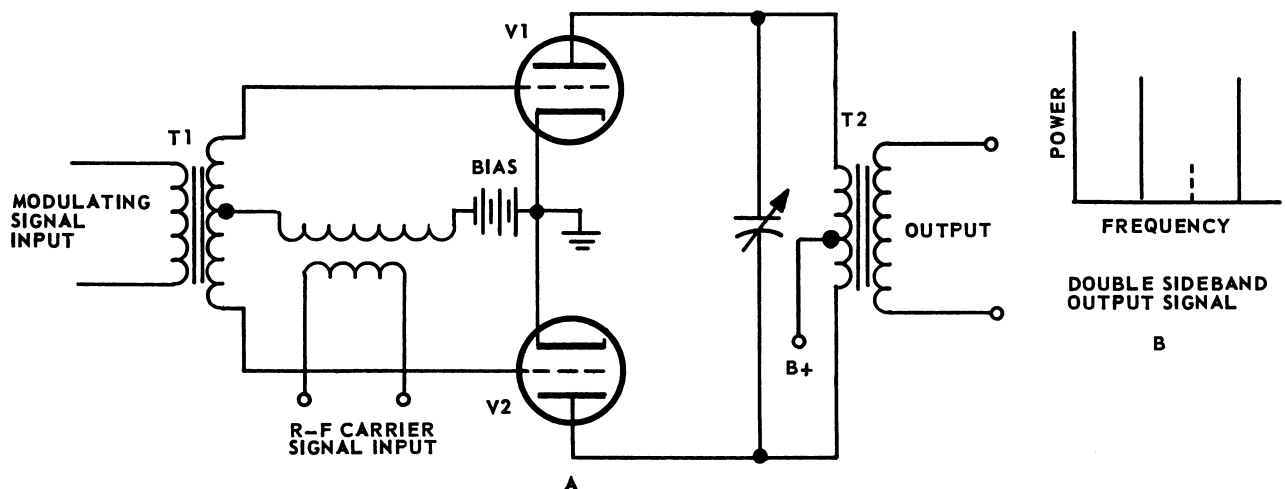


**Figure 9-33. Simplified Push-Pull Balanced Modulator**

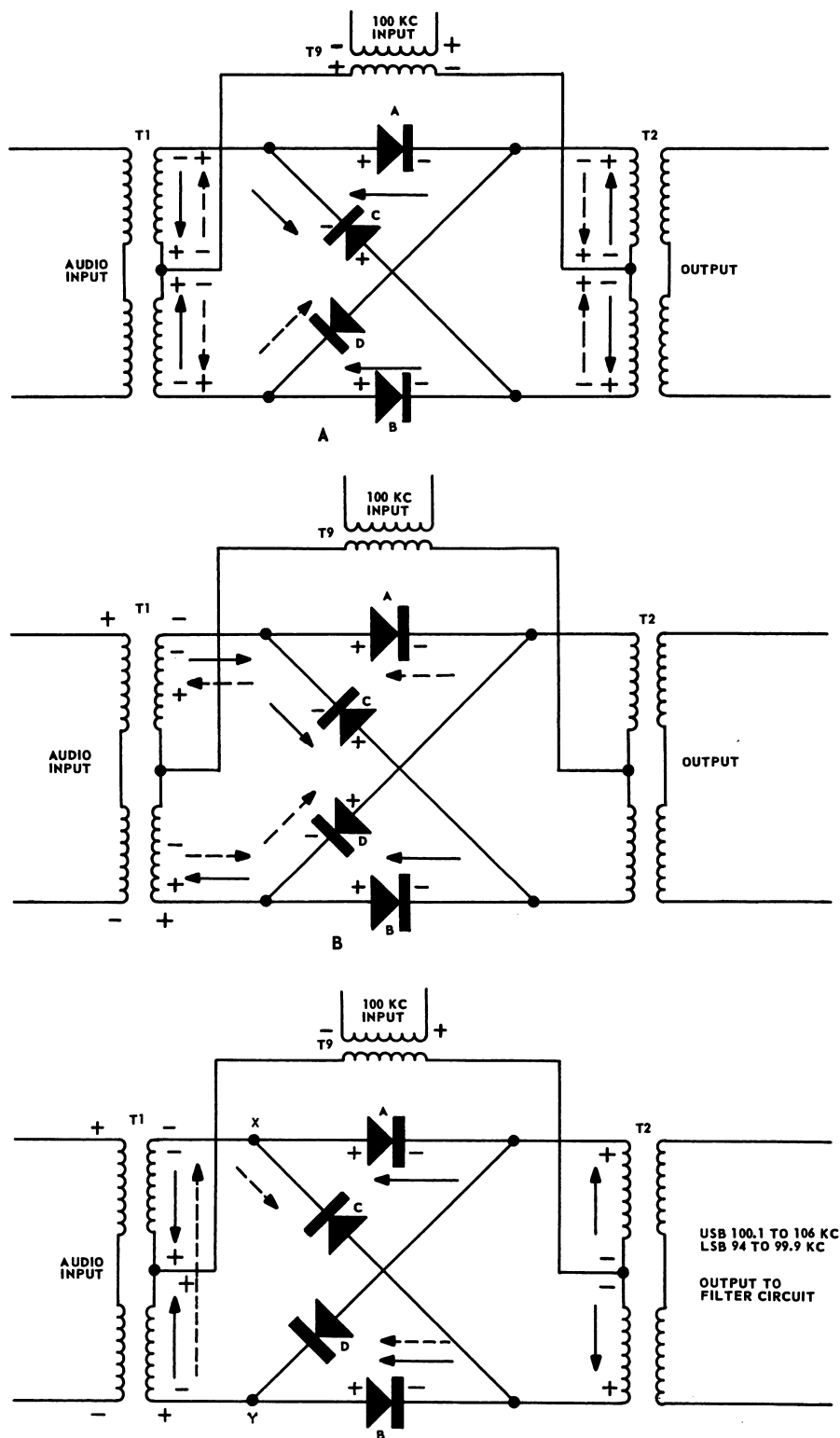


Figure 9-34. Balanced Ring Modulator

ing frequency, and the sum and difference frequencies flow through  $V_1$ . The carrier frequency currents are effectively cancelled by the carrier frequency currents from  $V_2$  in the common plate load as explained before. The sum and difference frequencies cause induced voltages in the secondary of transformer  $T_2$ , and the modulating frequency is so low that it has no appreciable effect on the plate coil.

The output signal of the balanced modulator consists of the upper and lower sidebands, the carrier frequency has been eliminated. This is shown in Figure 9-33B. Now that the carrier has been eliminated, all that remains is to suppress one sideband. This can be accomplished by either the filter method or the phasing method.

The balanced ring modulator shown in Figure 9-34 is also commonly used in Air Force SSB equipment. This modulator consists of four copper oxide varistors, which exhibit the property of variable resistance. The amount of resistance depends upon the magnitude and polarity of the voltage across the varistor. Electrons flow from the copper (bar) to the copper oxide (arrow) when the bar is negative with respect to the arrow; the varistor is said to be forward biased. When the bar is positive with respect to the arrow, there is no electron flow; the varistor is said to be back biased. The operation is that of the copper oxide rectifier which was explained in Chapter 4. The conduction through the varistors, therefore, is unidirectional with the magnitude of conduction depending upon the potential difference between the copper and the oxide.

To understand the operation of the ring-type balanced modulator, assume that a 100-kc r-f signal is applied to the circuit as in Figure 9-34A, with polarity across  $T_9$  as indicated. Also assume that there is no audio signal input. At this time the voltage across varistors A and B will allow conduction, while the voltage across varistors C and D prevents conduction. C and D act as an open circuit. Current flow is indicated by the solid arrows. The polarity of the r-f voltage developed across the secondary winding of  $T_1$  and the primary winding of  $T_2$  is indicated by the negative and positive signs at the ends of the solid arrows. The voltages developed across each section (center tap to either end) of the secondary of  $T_1$  are equal in amplitude and  $180^\circ$  out of phase. This also holds true for the r-f voltage developed across each section

(center tap to either end) of the primary winding of  $T_2$ .

Since voltages of equal amplitude and opposite polarity cancel, there will be no output voltage developed across the secondary winding of  $T_2$ . When the 100-kc input signal reverses polarity, the current will be in the direction indicated by the dotted arrows. The polarity of the voltage developed across the secondary winding of  $T_1$  and the primary winding of  $T_2$  is indicated by the signs at the ends of the dotted arrows. Now varistors C and D conduct and A and B do not conduct. The voltages developed across each section of the primary of  $T_2$  are again equal in amplitude and  $180^\circ$  out of phase. Therefore, with only a 100-kc r-f input, the circuit is balanced, and there is no output.

Figure 9-34B shows the same modulator with an audio-frequency input, but no 100-kc r-f input. When the polarity of the audio signal across the secondary winding is negative at the top, the voltages across the varistors are such that C and B conduct, while A and D do not conduct. The current path is indicated by the solid arrows. Current does not flow in the top half of the primary winding of  $T_2$  since A is effectively an open circuit. Varistors C and B are in series across the secondary of  $T_1$ . The secondary of  $T_9$  effectively connects the center taps of the  $T_1$  secondary and the  $T_2$  primary. Thus the bottom half of the  $T_2$  primary is connected between the center tap on the  $T_1$  secondary and the connection between varistors C and B. The ends of this section of the  $T_2$  primary are at the same potential in this network; therefore, no current flows in  $T_2$ .

When the polarity of the audio-frequency input reverses, the current path will be as indicated by the dotted arrows. Under these conditions, varistors A and D conduct, while C and B do not conduct. Current does not flow in the primary of  $T_2$  under these conditions. Thus, there is no output from the circuit when only an audio-frequency voltage is applied through transformer  $T_1$ .

Figure 9-34C shows the balanced ring modulator with both a-f and r-f inputs. Assume that the 100-kc r-f signal has the polarity indicated across  $T_9$  and that the audio signal (100 to 6000 cycles) has the polarity indicated across  $T_1$ . The amplitude of the r-f voltage is much greater than that of the audio voltage. With the polarity in-



dedicated for the r-f voltage, varistors A and B conduct in the direction shown by the solid arrows. The dotted arrows indicate the current due to the a-f voltage alone. Remember that the relative amplitudes and polarities of the two input signals are an important consideration in the analysis of the operation of this modulator. The actual voltages at points X and Y are the vector sums of the r-f and a-f voltages at the point concerned. At point X at the instant shown, the two voltages are opposing, while at point Y they are aiding.

The actual current flow through varistors A and B depends upon the summation of the applied voltages. The voltage developed across the top section of the secondary winding of T1 is the difference between the a-f and r-f voltages; the voltage developed across the bottom half of the T1 secondary is the sum of the two signal voltages. The voltage developed across the top section of the T2 primary is the difference between the two signal voltages, while that developed across the bottom section is the sum of the two voltages. The output voltage developed across the secondary of T2 is the resultant of the two voltages developed across the two sections of the primary of T2.

When the r-f voltage reverses polarity, varistors C and D conduct. The voltages at points X and Y are the resultants of the a-f and r-f voltages. The voltage developed across the top section of the T1 secondary is the sum of the two signal voltages, while that across the bottom section is the difference between the two signals. The voltages across the two sections of the T2 primary are the reverse of the voltages on the first r-f alternation. The output voltage across the secondary of T2 is the resultant of the voltages developed across the two sections of the T2 primary.

During the next alternation of the audio-frequency voltage, the top of T1 will be positive and the dotted arrows will be reversed. When the polarity of the r-f voltage is as shown across T9 in Figure 9-34C, the solid arrows again indicate conduction through varistors A and B. The voltage developed across the top half of the T1 secondary is now the sum of the two voltages, and the voltage across the bottom section of the T1 secondary is the difference between the a-f and r-f voltages. The voltage across the top section of the T2 primary is the sum of the two

voltages, and the voltage across the bottom section is the difference between the two signal voltages. Once again the output voltage across the secondary of T2 is the resultant of the voltages developed across the two sections of the T2 primary.

When the r-f voltage reverses polarity, varistors C and D conduct; the voltages subtract in the top section of the T1 secondary and add in the bottom section. This means that the difference voltage appears across the top section of the T2 primary, and the sum voltage across the bottom section. The output voltage is the resultant of the voltages developed across the two sections of the T2 primary.

Because the varistors are nonlinear devices, mixing of the two input signals in these elements produces the upper and lower sidebands. Since the ring-type modulator circuit is not balanced for these sideband frequencies, they will appear across the output transformer. Therefore, the output consists of the voltages developed from the sideband frequencies. The upper sideband frequencies extend from 100.1 kc to 106 kc, and the lower sideband frequencies extend from 94 kc to 99.9 kc. For all practical purposes, the 100-kc carrier signal and the audio-frequency modulating signal are balanced out in the modulator and do not appear in the output. Only the sideband frequencies containing the intelligence will appear in the output.

Both SSB transmitters and receivers require extremely sensitive bandpass filters in the region of 100 to 500 kc. In receivers a high order of adjacent channel rejection is required if channels are to be closely spaced in order to conserve spectrum space. In SSB transmitters the signal bandwidth must be limited sharply in order to pass the desired sideband and reject the other sideband. The filter used, therefore, must have very steep skirt characteristics (fast cutoff) and a flat bandpass characteristic in order to pass all frequencies in the band equally well. These filter requirements are met by crystal filters, LC filters, and mechanical filters.

### Simple Filters

Filters are designed to produce high attenuation at particular frequencies. Filters are made up of inductors and capacitors, arranged to offer a high loss to currents of undesired frequencies and a low loss to other frequencies. There are

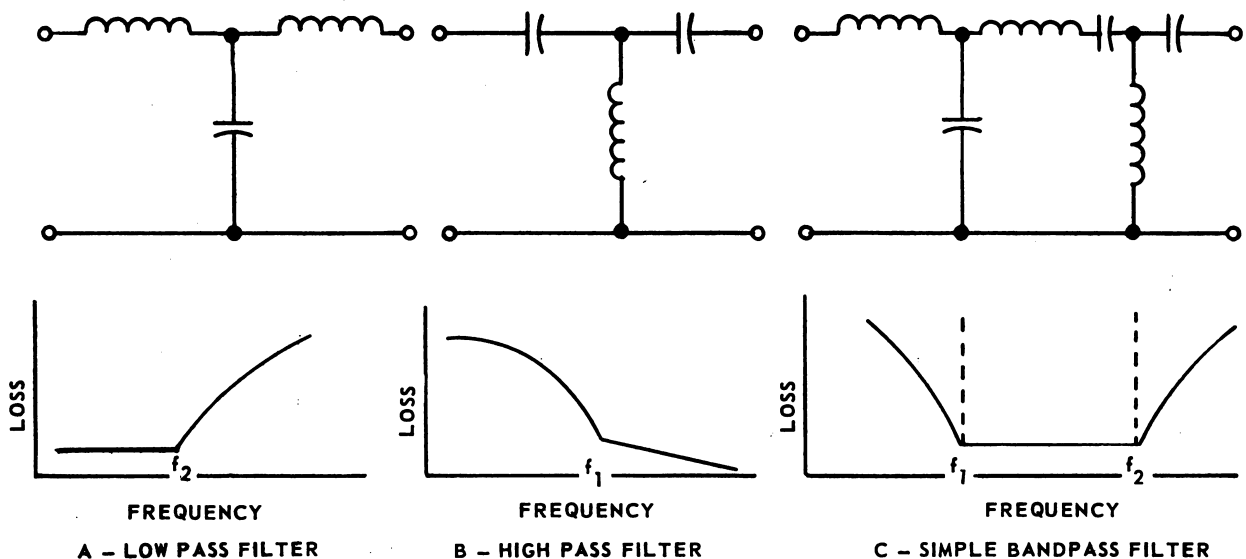


Figure 9-35. Basic Filters

three types of filters: the low-pass filter, the high-pass filter, and the bandpass filter.

A simple low-pass filter is shown in Figure 9-35A. When low-frequency currents are applied to a low-pass filter, the filter causes very little loss. At low frequencies the reactance of the series inductors is low and the reactance of the shunt capacitor is high. However, when high-frequency currents are applied to a low-pass filter, the reverse is true; the reactance of the series inductors is high, and the reactance of the shunt capacitor is low. The frequency response of a low-pass filter is illustrated below the schematic diagram;  $f_2$  is the highest frequency passed.

In a high-pass filter the capacitors are in the series arm and the inductor is in the shunt arm. A high-pass filter offers a high loss to low-frequency and a low loss to high-frequency currents. The relation between loss and frequency is shown in Figure 9-35B. In this illustration,  $f_1$  is the lowest frequency passed.

A simple bandpass filter can be made from a combination of these two low- and high-pass filters, as shown in Figure 9-35C. Assume that the low pass section (coils in the series arm) is designed to pass a frequency range from zero to 2000 cycles and to suppress frequencies above that point. In addition, assume that the high pass section (series capacitors) is designed to pass a frequency range from 1000 cycles up.

These two filters in series, then, will pass only frequencies between 1000 and 2000 cycles. These frequencies mark the lower and upper limits of the bandpass. This is shown in the frequency response curve of the simple bandpass filter in Figure 9-35C. None of the three filter circuits shown would actually eliminate the unwanted sideband, although the simple bandpass filter comes the nearest to doing so.

#### Bandstop Filter

From your study of resonant circuits you know that a series resonant circuit offers minimum opposition to the resonant frequency. A circuit such as that shown in Figure 9-36A would effectively short out the frequency to which the series circuit was resonant. All other frequencies would be passed down the line. A parallel resonant circuit offers maximum opposition to the frequency of resonance. As shown in Figure 9-36B, the parallel resonant circuit in the line offers high opposition to the frequency to which the tank is resonant, and low impedance to other frequencies.

A filter made up of both series and parallel resonant circuits, such as that shown in Figure 9-36C, would prove effective in eliminating one sideband. Let us assume that both sidebands are applied to the input of this bandstop filter, one of 795 kc and the other of 805 kc. The

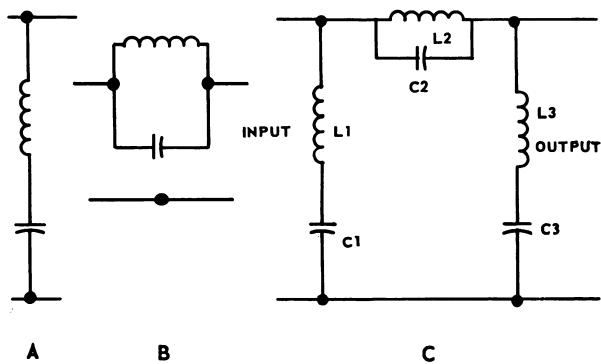


Figure 9-36. Bandstop Filter

balanced modulator has already eliminated the carrier frequency of 800 kc. We wish to eliminate the upper sideband (805 kc).  $L_1C_1$  and  $L_3C_3$  form two series resonant circuits whose resonant frequency is 805 kc.  $L_2C_2$  is a parallel resonant circuit whose frequency of resonance is also 805 kc. When the two sideband frequencies are applied to this filter,  $L_1C_1$  effectively bypasses the 805-kc signal to ground, and develops a high voltage for the 795-kc signal. The parallel resonant circuit of  $L_2C_2$  in the line offers maximum opposition to 805 kc, effectively blocking it.

Of course, it is extremely difficult to construct a resonant circuit with sharp cutoff frequency characteristics, so that a small amount of the 805-kc signal will not be eliminated. However, any of the 805-kc signal passing the parallel resonant circuit will be eliminated by the series resonant combination of  $L_3C_3$ , which offers minimum impedance to that frequency. Now, only the lower sideband, 795 kc, will be developed across this filter circuit; the upper sideband has been attenuated to the point where it is no longer of any effect.

### Crystal Filter

The crystal filter may also be used in single sideband systems to attenuate the unwanted sideband. This is possible since the crystal is a series resonant circuit. Because of its very high  $Q$ , the crystal filter passes a much narrower band of frequencies than the best LC filter. Because of its high selectivity it reduces the number of undesired frequencies in the output.

The equivalent circuit of the crystal and crystal holder is illustrated in Figure 9-37A. The components  $L$ ,  $C_1$  and  $R$  represent the series

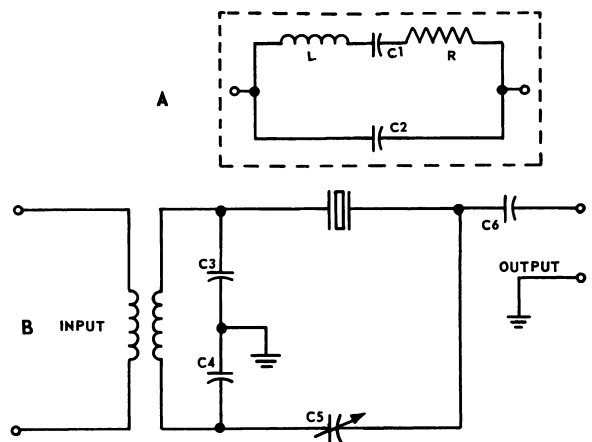


Figure 9-37. Crystal Filter Circuit

resonant circuit of the crystal itself.  $C_2$  represents the parallel capacitance of the crystal holder. The crystal offers a very low impedance path to the frequency to which it is resonant and a high impedance path to other frequencies. However, the crystal holder capacitance,  $C_2$ , shunts the crystal and offers a path to other frequencies. For the crystal to operate as a bandpass filter, some means must be provided to counteract the shunting effect of the crystal holder. This is accomplished by placing an external variable capacitor in the circuit.

In Figure 9-37B variable capacitor  $C_5$ , called the *phasing capacitor*, counteracts holder capacitance  $C_2$ .  $C_5$  can be adjusted so that its capacitance equals the capacitance of  $C_2$ . Then both  $C_2$  and  $C_5$  pass undesired frequencies equally well. Because of the circuit arrangement, the voltages across  $C_2$  and  $C_5$  due to undesired frequencies are equal and  $180^\circ$  out of phase. Therefore, undesirable voltages cancel, and undesired signals will not appear in the output. Actually, there is only one frequency in the filter's passband at which total cancellation can be obtained.  $C_5$  can be adjusted to cause exact cancellation at any frequency within the passband except for frequencies very close to the crystal's resonant frequency. This cancellation effect is called the *rejection notch*.

To understand why the rejection notch may be positioned as desired by varying  $C_5$ , the effective impedance of the holder-crystal combination must be considered. At the crystal's series-resonant frequency, its capacitive and inductive reactances are equal and effectively cancel. As the

frequency increases, the inductive reactance increases while the capacitive reactance decreases so that the crystal's impedance becomes inductive. At some frequency slightly higher than the crystal's series-resonant frequency, this net inductive reactance of the crystal equals the capacitive reactance of  $C_2$ . This is the parallel resonant frequency of the holder-crystal combination. At still higher frequencies, the net inductive reactance of the crystal exceeds the capacitive reactance of  $C_2$  so that the impedance of the holder-crystal combination (parallel circuit) becomes capacitive.  $C_5$  can be adjusted to be equal to this effective capacitance causing cancellation at the output.

Thus, the rejection notch can be positioned at frequencies above the holder-crystal combination's parallel-resonant frequency. At frequencies below series resonance, the crystal's capacitive reactance exceeds its inductive reactance and the net crystal impedance is capacitive. This capacitance, paralleled with  $C_2$ , causes the holder combination's impedance to be capacitive.  $C_5$  can be adjusted to equal this value of capacitance and cause cancellation at the output.

The effective impedance of the holder-crystal combination is capacitive at frequencies both below the crystal's series-resonant frequency and above the holder-crystal combination's parallel-resonant frequencies.  $C_5$  can be adjusted, then, to equal the effective capacitance of the holder-crystal combination and to position the rejection notch at the desired point.

For circuit operation, assume that a lower sideband frequency of 795 kc and an upper side-

band frequency of 805 kc are applied to the input of the crystal filter. As in the previous example, assume that the upper sideband is the unwanted sideband. By selecting a crystal that will provide a low impedance path at 795 kc, the lower sideband frequency will appear in the output. The upper sideband, as well as all other frequencies, will have been attenuated by the crystal filter.

### Mechanical Filters

Mechanical filters are now used in single sideband equipment because of recent advancements in the development of this type of filter. Some of the advantages of mechanical filters are their excellent rejection characteristics, extreme ruggedness, size small enough to be compatible with the miniaturization of equipment, and a  $Q$  in the order of 10,000, which is about 100 times that obtainable with electrical filters.

The basic principles of mechanical filters are well established, although their commercial use is relatively new. The mechanical filter is a device that is mechanically resonant; it receives electrical energy, converts it to mechanical vibration, then converts this mechanical energy back into electrical energy at the output. There are four elements constituting a mechanical filter: (1) an input transducer that converts the electrical energy at the input into mechanical vibrations, (2) metal disks that are manufactured to be mechanically resonant at the desired frequency, (3) rods that couple the metal disks, and (4) an output transducer that converts the mechanical vibrations back into electrical energy.

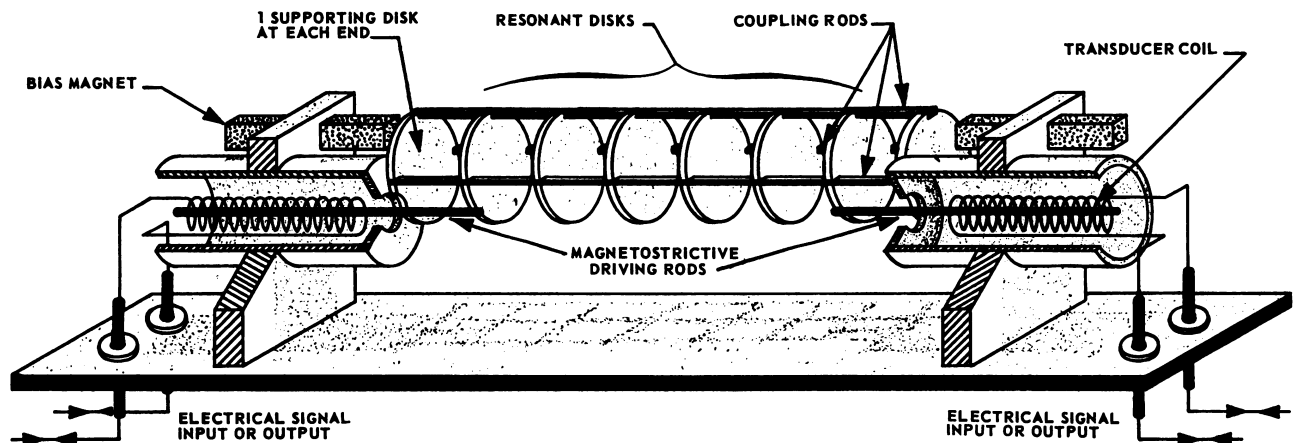


Figure 9-38. Mechanical Filter

Figure 9-38 shows the elements of a mechanical filter. Not all of the disks are drawn in the illustration. The shields around the transducer coils have been cut away to show the coil and magnetostrictive driving rods. As you can see, either end of the filter may be used as the input.

Figure 9-39 is the electrical equivalent of the mechanical filter. The disks of the mechanical filter are represented by the series resonant circuits  $L_1C_1$  while  $C_2$  represents the coupling rods.  $R$  in both input and output represents the matching mechanical loads.

**TRANSDUCERS.** The transducer converts mechanical energy into electrical energy or electrical energy into mechanical energy. Both magnetostrictive and electrostrictive devices are transducers. Electrostriction is defined as the deformation produced by electric stress when a dielectric is charged. Since electrostriction in dielectrics applies to any interaction between an electric field and the corresponding deformation, electrostriction includes the piezoelectric effect. An electrostrictive transducer such as a piezoelectric crystal will compress when subjected to an electric current or voltage.

The magnetostrictive transducer depends on the principle that certain materials can be made to lengthen or shorten when in a magnetic field. If an electrical signal is sent through a coil whose core is made of a magnetostrictive material, the electrical variations will be converted into mechanical changes. The mechanical oscillations can be used to drive the mechanical elements of a filter, which are the disks.

Both types of transducers have a natural resonant frequency. They are manufactured or tuned to be resonant at a specific frequency. Transducers are chosen for a particular application, in this instance, the ability to pass the frequencies

of one sideband and to reject all other frequencies.

**OPERATION.** Let us assume that the mechanical filter of Figure 9-38 has disks tuned to pass the frequencies of the desired sideband. The input to the filter contains both sidebands, and the transducer driving rod applies both sidebands to the first disk. The vibration of the disk will be greater at the frequency to which it is tuned (resonant frequency), which is the desired sideband, than at the undesired sideband frequency. The mechanical vibration of the first disk is transferred to the second disk, but a smaller percentage of the unwanted sideband frequency is transferred. Each time the vibrations are transferred from one disk to the next, there is a smaller amount of the unwanted sideband. At the end of the filter there is practically none of the undesired sideband left. The desired sideband frequencies are taken off the transducer coil at the output end of the filter.

Varying the size of  $C_2$  in the equivalent circuit in Figure 9-39 varies the bandwidth of the filter. Similarly, by varying the mechanical coupling between the disks (Figure 9-38), that is, by making the coupling rods either larger or smaller, the bandwidth of the mechanical filter is varied. Because the bandwidth varies approximately as the total area of the coupling rods, the bandwidth of the mechanical filter can be increased by using either larger coupling rods or more coupling rods. Mechanical filters with bandwidths as narrow as 0.5-kc and as wide as 35 kc are practical in the 100- to 500-kc range.

### Single-Sideband Transmitters

Figure 9-40 is a block diagram of a single sideband suppressed carrier transmitter using me-

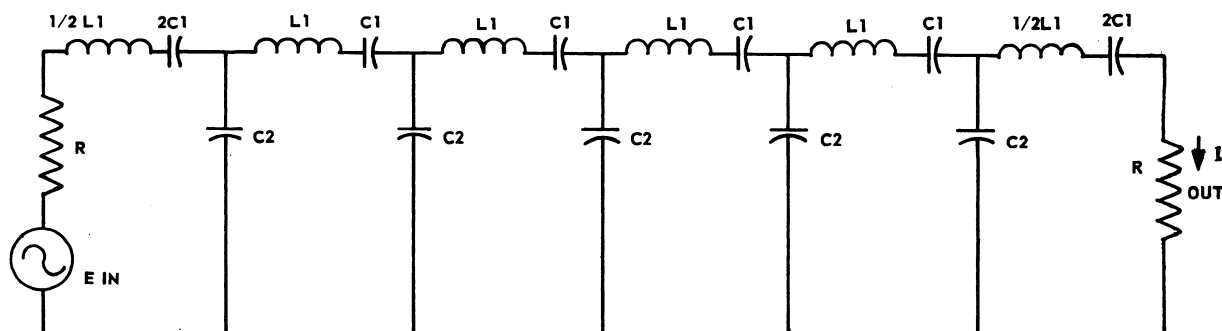


Figure 9-39. Electrical Analogy of a Mechanical Filter

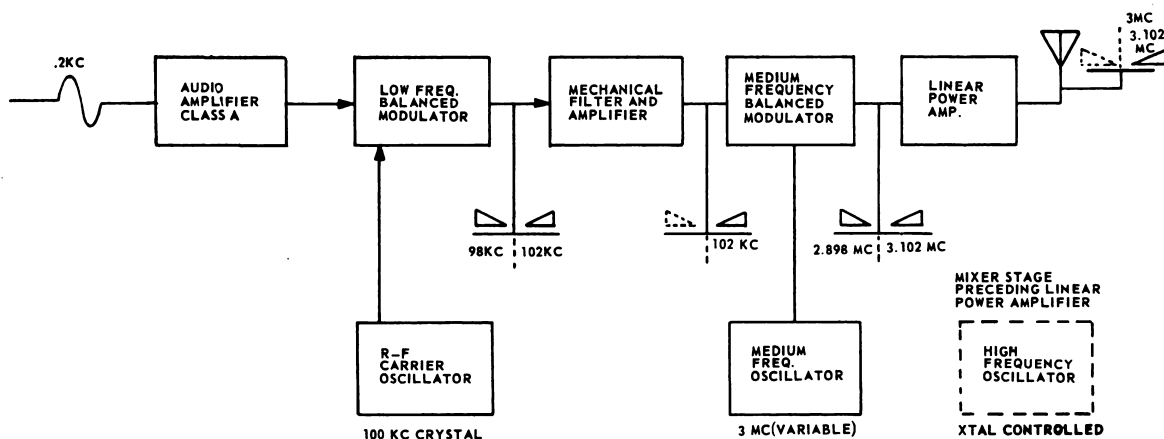


Figure 9-40. SSB Transmitter Block Diagram

chanical filters and a balanced modulator. For illustrative purposes, a single tone, 2000-cycle audio signal is used, but in actual practice a complex audio signal would be applied.

The 2-kc signal is amplified and mixed with a 100-kc carrier or conversion frequency in the balanced modulator. Remember, neither the carrier nor the audio frequency appear in the output of the balanced modulator; but the sum and difference frequencies (98 kc and 102 kc) do appear in the output. As illustrated in Figure 9-40, two sidebands are applied to the mechanical filter. Only the desired upper sideband is passed. The dotted lines show where the carrier and lower sideband have been removed.

Because the remaining sideband (containing the intelligence) is too low in frequency to transmit efficiently, it must be mixed again with a new conversion frequency to raise it to the desired transmitter frequency. The 3-mc oscillator applies a signal to a balanced modulator. Again, the balanced modulator, after mixing the two inputs to get two new sidebands, removes the two fundamentals and applies the two new sidebands (3102 kc and 2898 kc) to a tunable linear power amplifier.

The input and output circuits of the linear power amplifier are tuned to reject one sideband and pass the other to the antenna for transmission. The high frequency oscillator is variable so that the transmitter output frequency can be varied over a range of operating frequencies. Since both the carrier and one sideband have been eliminated, all of the transmitted energy is in the single sideband.

Under certain applications another medium-frequency oscillator and modulator may be inserted for increased stability and a wider operating band.

The independent sideband type of SSB transmitter generates both sidebands, each containing different intelligence, with the carrier suppressed to a desired level. Basically, this is accomplished with a triple conversion process that translates the two audio frequency inputs (100 to 6000 cps) into two r-f sideband signals centered around a reinserted pilot carrier. The pilot carrier is used as a reference and to control receiving equipment auxiliary circuits such as AVC, AFC, and so forth. Figure 9-41 illustrates a typical block diagram for an independent sideband transmitter.

The first conversion step converts the audio inputs into a low frequency (LF) signal of  $100 \pm 6$  kc. Each audio input is applied to a separate balanced modulator at the same time as a 100-kc conversion frequency. The output of each modulator will contain the upper sideband frequency of 100.1 to 106 kc and the lower sideband frequency of 94 to 99.9 kc. The 100-kc conversion frequency and the audio inputs will not appear in the modulator output.

Associated with each modulator is a lattice type crystal filter. The Group A signal is fed to the A crystal filter, which passes only the upper sideband (sum frequency) output from LF Mod 1. The Group B signal is fed to the B filter, which passes only the lower sideband (difference frequency) output from LF Mod 2. At this point we have succeeded in generating, or developing, the independent sidebands.

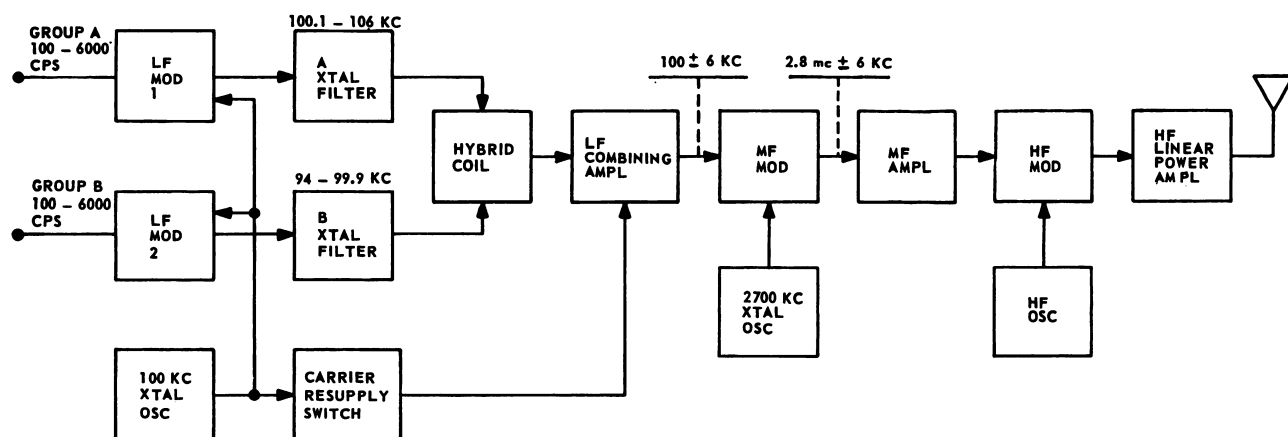


Figure 9-41. Independent Sideband SSB Transmitter Block Diagram

This brings up another problem. How can these two signals be transmitted? To accomplish this task, we use what is termed a *hybrid coil*. This is a specially designed transformer that combines two signals without mixing or interaction. The hybrid coil is a linear device which combines the two sideband signals into one composite signal voltage that represents the two bands of frequencies of 94 to 99.9 kc and 100.1 to 106 kc. The output of the hybrid coil is then applied to the LF combining amplifier.

The 100-kc crystal-controlled oscillator supplies the conversion frequency for the LF modulators and the signal that will be used as the reinserted carrier. This signal is applied to the carrier resupply switch, which is adjusted for any desired level of carrier suppression. Normal independent sideband operation requires that the carrier be suppressed 20 db with respect to the level of the sidebands of intelligence. When the carrier of a transmitter rated at 4000 watts peak envelope power is suppressed 20 db, there will be approximately 40 watts of power in the carrier. The remaining 3960 watts of power is contained in the upper and lower sidebands.

The LF combining amplifier in Figure 9-41 is a Class A linear amplifier. Its function is to combine the sidebands with the reinserted carrier. The output of this amplifier is applied to the medium-frequency (MF) modulator, where the second frequency conversion takes place. The output of the MF modulator contains the  $2600 \pm 6$  kc lower sideband and the  $2800 \pm 6$  kc upper sideband. The LF input signal and 2700-kc conversion frequencies are balanced out in the modulator circuit. The undesired modula-

tion products are passed to ground by resonant filters, while the sum frequency is passed by a filter in the MF amplifier input. The  $2800 \pm 6$ -kc signal is amplified by the MF amplifier and then applied to the next conversion stage.

The final frequency conversion takes place in the HF modulator when the MF signal is mixed with the conversion frequency generated by the HF oscillator, which determines the operating frequency of the transmitter. This oscillator makes possible an output over a wide range of frequencies in the high-frequency spectrum. Assuming that the transmitter is to operate on 5 mc, the HF oscillator must produce an output of 7.8 mc or 2.2 mc, depending on whether the HF linear power amplifier is tuned to the upper or the lower sideband output of the HF modulator circuit. To operate on 25 mc, the HF oscillator must produce an output of either 22.2 mc or 27.8 mc, depending on how the linear power amplifier is tuned. The HF linear amplifier is always tuned to the operating frequency, which will be either the upper sideband (sum) or the lower sideband (difference) output of the HF modulator.

### Single-Sideband Receivers

To see the relationship of the parts in a single sideband receiver, observe the block diagram in Figure 9-42. Basically, the receiver is similar to an ordinary AM superheterodyne; that is, it has r-f and i-f amplifiers, mixer, detector, and audio amplifiers. However, to permit satisfactory SSB reception, an additional mixer (demodulator) and oscillator must replace the conventional diode second detector.





put. Consider an SSB of 2 mc to be transmitted. This signal is generated by combining a carrier of 1997 kc with an audio frequency of 3 kc in a balanced modulator and selecting the upper sideband,  $1997 \text{ kc} + 3 \text{ kc} = 2 \text{ mc}$  (2000 kc). When the receiver is tuned to this frequency, the 2-mc signal is converted to a 455-kc signal, amplified and applied to the second mixer. When the carrier oscillator is adjusted to 458 kc, a 3-kc signal (458 kc - 455 kc) is produced in the output of the mixer.

Tuning the sideband receiver is somewhat more difficult than a regular AM receiver. The carrier (beat frequency) oscillator must be precisely adjusted to simulate the carrier frequency at all times. Any tendency of drift within the oscillator will cause the output intelligence to be distorted.

In order to separate the sidebands when both are transmitted, a selective crystal filter is placed in the i-f amplifier strip. This filter is capable of shifting the passband slightly above or below the normal i-f frequency for reception of the upper or lower sideband respectively.

The independent sideband type of SSB receiver (Figure 9-43) is basically the same as the SSSC type except that this receiver must have two audio outputs. This is necessary to reproduce the two audio signals originally applied to the transmitter. The basic principles involved in this type of reception are the same as those previously explained for the transmitter.

The first and second conversion stages of the independent sideband receiver are almost identical to those used by the normal military com-

munications receiver operating on the super-heterodyne principle. The main difference in this receiver is the use of a hybrid coil that separates the  $100 \pm 6 \text{ kc}$  signal into two signals before application to their respective filters and demodulators. The A crystal filter passes frequencies from 100.1 to 106 kc, while the B filter passes the lower sideband frequencies from 94 to 99.9 kc.

The signals are each demodulated by their own separate balanced demodulator when mixed with the 100-kc conversion frequency. These demodulators are identical to the balanced ring modulator. The two 100- to 6000-cycle audio signals are amplified by their respective amplifiers and then applied to the external terminal equipment. The signal that is taken from the LF amplifier is used to operate the special circuits required for AVC, AFC, and reconditioned carrier applications.

#### Pre-Emphasis and De-Emphasis Networks

Audio frequency pre-emphasis and de-emphasis networks are also known as audio frequency accentuation and deaccentuation circuits, respectively. Where the former network is used in the FM transmitter, the latter network will be used in the FM receiver.

The amplitudes of the higher audio frequencies in the audio input to a transmitter are comparatively low when considering the whole audio frequency spectrum. This is not due to any transmitter characteristic, but to the natural distribution of sound in radio program material and communications as used by the military. Noise is unavoidably added as the transmission

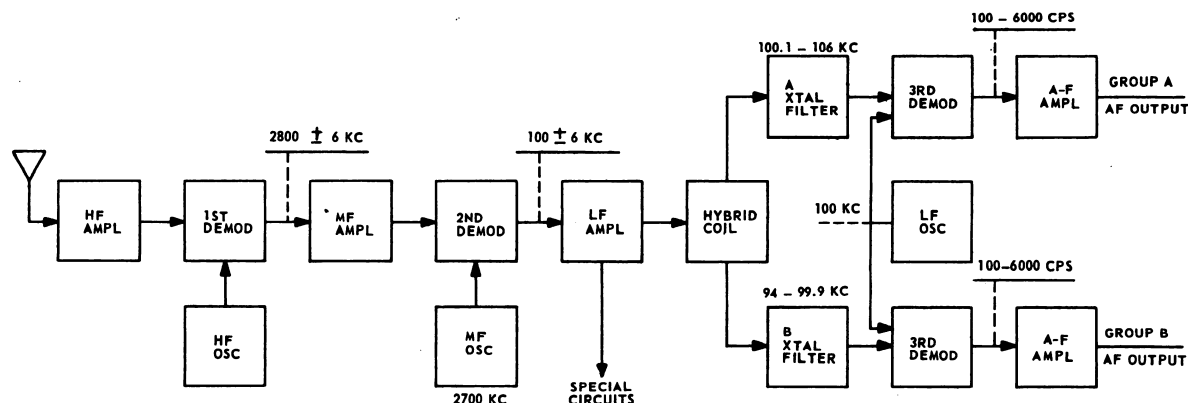


Figure 9-43. Independent Sideband SSB Receiver Block Diagram

passes through the transmitter and then on through the receiver.

The useful relative amplitudes of the desired audio frequency signals are generally small in the upper audio frequency range. Since noise is predominantly high-frequency audio, this condition tends to produce an undesirable signal-to-noise ratio at the high audio frequencies. The signal is relatively weak and the noise is relatively strong. Pre-emphasis circuits are used in the transmitter to overcome this condition.

By definition, pre-emphasis is the increasing of the relative strength of the high-frequency components of the audio signal before it is fed to the modulator. Stated differently, pre-emphasis provides in the upper audio-frequency range the desired signal currents, with respect to desired signal currents in the lower audio-frequency range. Thus, the undesirable relationship between the high-frequency signal and the high-frequency noise is altered. While the noise remains the same, the signal strength is increased. There is a disadvantage, however, in the change in the natural balance between high- and low-frequency tones.

A de-emphasis circuit in the receiver, however, corrects this defect, as it reduces the high-frequency audio the same amount as the pre-emphasis circuit increased it, thus regaining the original tonal balance. In addition, the de-emphasis network operates on both the high-frequency signal and the high-frequency noise; therefore, there is no change in the improved signal-to-noise ratio.

The main reason for the pre-emphasis network then, is to prevent the high-frequency components of the transmitted intelligence from being blocked out by the high-frequency noise characteristics inherent in the transmitter. In the intelligence this same pre-emphasis characteristic will be picked up by the receiver.

The de-emphasis network is normally inserted between the detector and the audio amplifier. This makes certain that the audio frequencies are returned to their original relative level before amplification. The pre-emphasis characteristic curve is flat up to 500 cycles. From 500 to 15,000 cycles there is a sharp increase in gain up to approximately 17 db. The gain at these frequencies is necessary to maintain the high signal-to-noise ratio at high audio frequencies. The frequency characteristic of the de-emphasis

network is directly opposite to that of the pre-emphasis network. The high-frequency response decreases in proportion to its increase in the pre-emphasis network. The characteristic curve of the de-emphasis circuit should be a mirror image of the pre-emphasis characteristic curve.

If the two characteristic curves are mirror images of each other, the time constant of both networks should be the same. For commercial broadcasting a time constant of 75 microseconds has been set as a standard. For military pur-

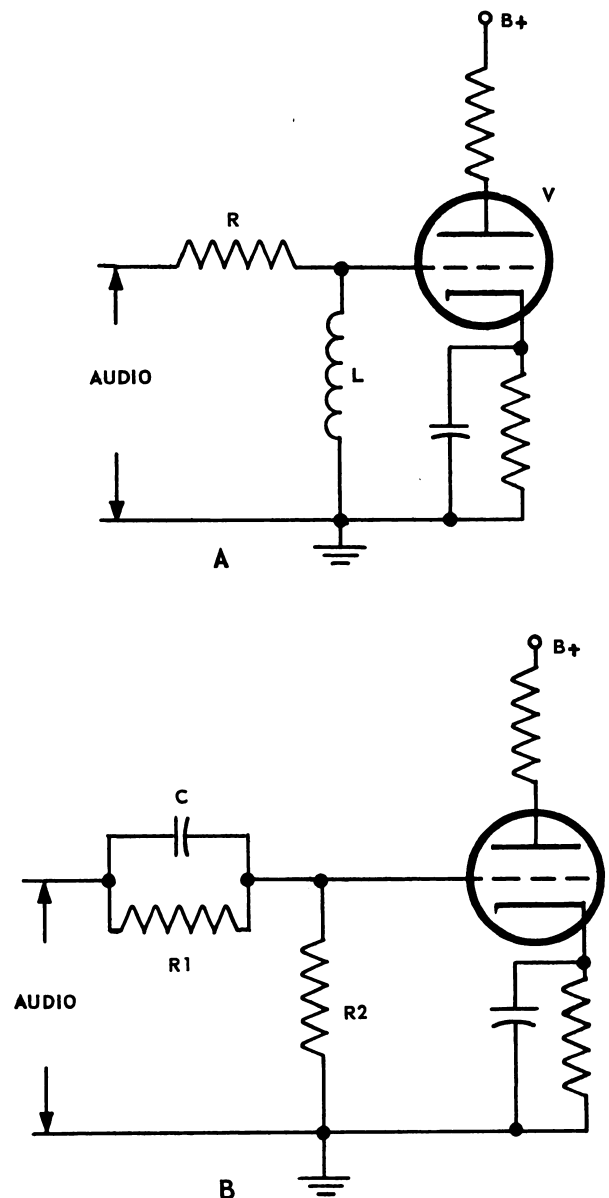


Figure 9-44. Typical Pre-Emphasis Circuits

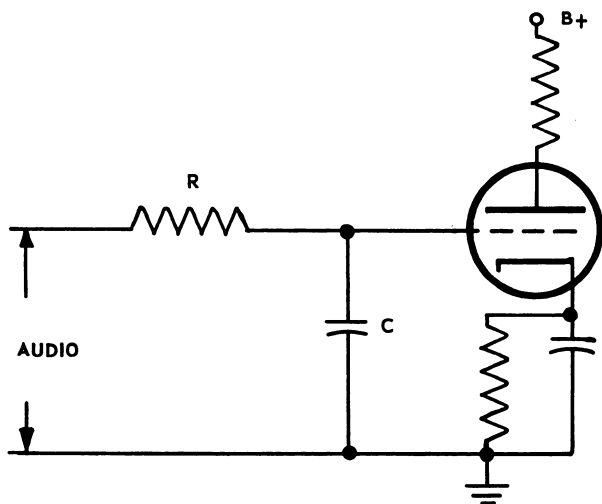


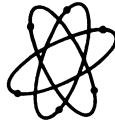
Figure 9-45. De-Emphasis Circuit

poses design specifications will determine the time constant.

Two typical pre-emphasis networks are illustrated in Figure 9-44. In Figure 9-44A the series circuit of  $R$  and  $L$  forms the total impedance to the audio. Since  $X_L = 2\pi fL$ , if  $f$  (frequency) increases, then  $X_L$  increases. This condition

being true, an increase in frequency will produce an increase in voltage across  $L$ . This is the input to the grid of the tube. In Figure 9-44B the impedance to the audio voltage is that of the parallel circuit of  $C$  and  $R_1$ , as the ohmic value of  $R_2$  is small in comparison to that of either  $C$  or  $R_1$ .  $X_c = \frac{1}{2\pi fC}$ . As the audio frequency increases, the reactance of  $C$  decreases. This decrease of  $X_c$  provides an easier path for high frequencies as compared to  $R$ . Thus, with an increase of audio frequency, there is a like increase in signal current. The result is a larger voltage drop across  $R_2$  at the higher frequencies.

Figure 9-45 depicts a typical de-emphasis network. Note the physical position of  $R$  and  $C$  in relation to the grid of the tube. As the frequency of the audio signal increases, the reactance of capacitor  $C$  decreases. The voltage division between  $R$  and  $C$  now provides a smaller drop across  $C$ . The audio voltage applied to the grid of the tube decreases; therefore, a reverse of the pre-emphasis circuit is accomplished. For the signal to be exactly the same as before pre-emphasis and de-emphasis, the time constants of the two circuits should be equal to each other.



# Transmission Lines

A transmission line is a conductor or a group of conductors used singly or in multiple to transmit electrical energy from a source to a load. Transmission lines are necessary because the load is normally located at a distance from the source of the energy. The primary purpose of a transmission line is to carry electrical energy from the source to the load with minimum loss.

This chapter is to acquaint you with the principles of transmission lines. It discusses their theory, uses, and construction. It also explains the types of losses to which radio frequency lines are subject, methods of making measurements on them, and how the principles of radio frequency lines are used in constructing artificial lines.

## DEFINITION

A radio frequency line is a special form of transmission line. A transmission line is a device, usually a pair of electrical conductors, that transfers electrical energy from a power source to a distant electrical circuit requiring power. The using device is called the *load*.

In transferring DC power the sole requirement of a transmission line is that it offers low DC resistance. In transferring AC power at frequencies as low as 60 or 400 cycles per second (commercial and aircraft power frequencies), the reactance of a transmission line is negligible compared to its resistance. As frequency increases, however, the effects of capacity and inductance increase, until at radio frequencies (above 50 kc) the reactive components retard the transfer

of energy to such an extent that a complete cycle may be generated at the source before the voltage at the start of the cycle reaches the load. Whenever this occurs, the line is said to be *electrically long*. At the frequency of 400 cps it takes a line about 400 miles long to cause this phenomenon. The longest line carrying 400-cycle frequencies possible in an aircraft is about 200 feet. Such a line is considered very short electrically.

In such cases it is possible to disregard the reactive effects. But at a frequency of 10,000 megacycles (a frequency commonly used in radar), a distance of only 2 inches in a line causes this effect. The 2-inch line is electrically long. A transmission line is considered to be electrically short when its physical length is small compared to a quarter-wavelength of the energy it is to carry. A line is said to be electrically long when its length is great in comparison to a quarter-wavelength. This means that at these high frequencies used in radar you must consider the length of the conductor from one vacuum tube to another as electrically long. Whenever a line is electrically long it is considered a *radio-frequency line*, and as such there are special considerations in design and adjustment.

## FUNCTION

In radar equipment r-f lines transfer power from the transmitter to the antenna and from the antenna to the receiver. Since the r-f line displays reactive characteristics, short sections

of r-f lines are used as tuned oscillator circuits, as inductors or capacitors, and as delay networks. The capacity in an r-f line can be charged, and upon discharging, produces a square wave rather than an exponential one. The r-f line is also used to produce a square radar transmitter pulse.

### R-F LINE THEORY

By resolving an r-f line into a simple equivalent circuit and analyzing this circuit, you will be able to understand the characteristics of the r-f line. This equivalent circuit is made by lumping all the resistances into a single large resistor, all the inductances into a single large inductor, and so on. Such a circuit is an r-f line with

lumped constants. To evaluate the effect of the r-f line shown in Figure 10-1A, it is necessary to put only the constants of a small section of a line into the equivalent circuit and to multiply the effect of this circuit by the number of these sections.

The first constant to consider is the DC resistance of the conductors. In the equivalent circuit this constant is a series-connected element. It is represented by  $R$  in each line of the equivalent circuit shown in Figure 10-1E.

The second constant is inductance. Previously you learn that whenever electrons move in a wire, a magnetic field is produced about the wire which, because of self-induction, retards the acceleration of the electrons as it builds up. In an r-f line there is a similar self-inductive action.

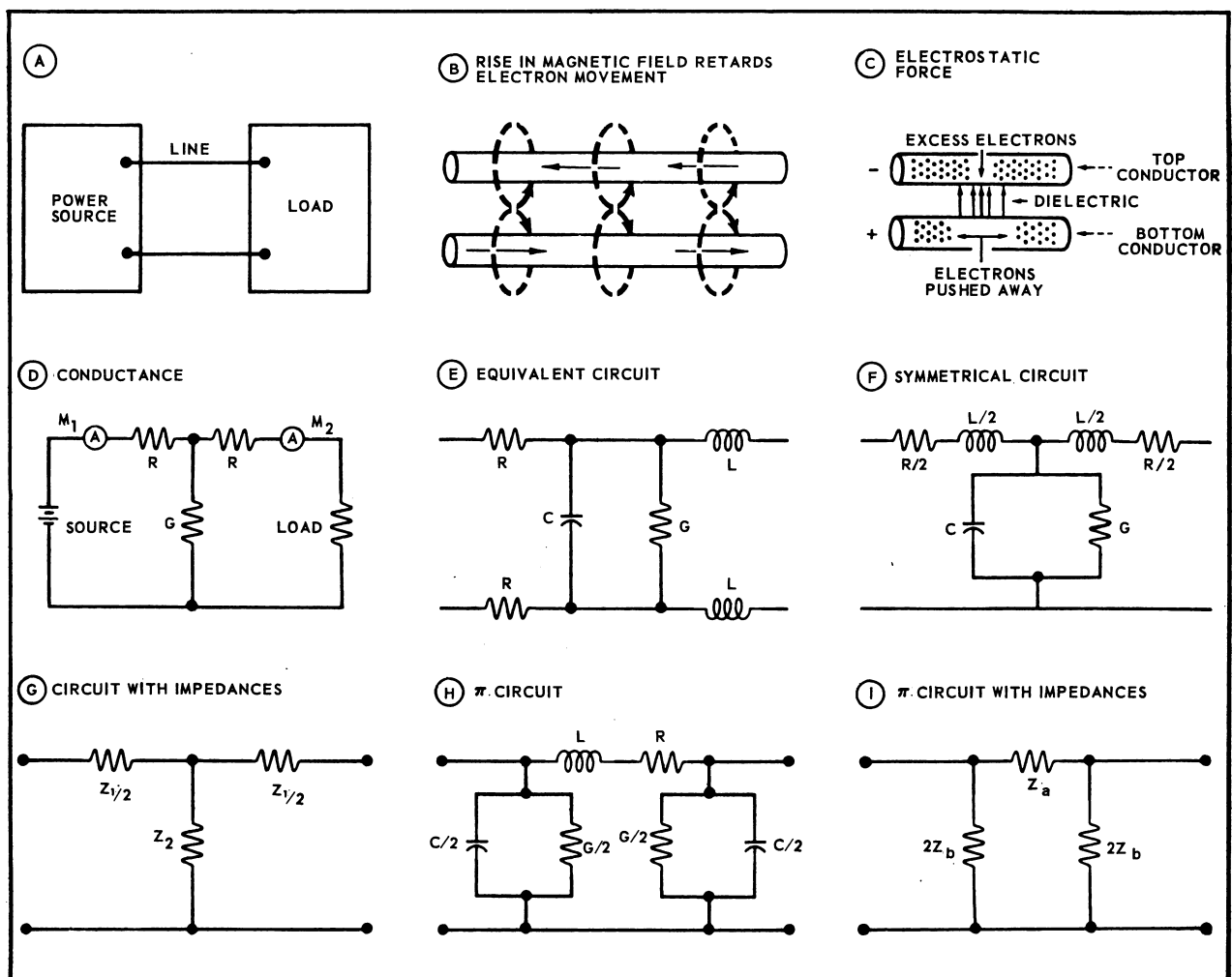


Figure 10-1. Equivalent Circuit of R-F Transmission Line

This action is represented in the equivalent circuit by inductors in each line, as shown in Figure 10-1.

Another constant to consider is capacitance. To understand the effect of capacitance in r-f lines, recall that the term *voltage difference* means that an excessive number of electrons exist at a certain point in an electrical circuit. In an r-f line, voltage difference means that there are more electrons on one conductor than on the other. The point of excess electrons exerts a force through the dielectric material between the two conductors and pushes the electrons away from this point as indicated in Figure 10-1C. This action is identical to the action that takes place between the two plates of a capacitor, and indicates that there is a capacitance between the two conductors. The capacitor  $C$  in the equivalent circuit at E represents the capacitance of the r-f line.

The fourth constant in an r-f line is the leakage that takes place from one wire to the other through the insulation between the wires. There is no such thing as a perfect insulator. Any insulator, therefore, will act as high resistance rather than an open circuit. The insulation conductance between wires is represented as  $G$  in Figure 10-1D. In this case note that with  $G$  removed from the equivalent circuit, the current on meter  $M_1$  at the source will read the same as meter  $M_2$  at the load. When you replace the shunt conductance ( $G$ ) in the circuit, more current will flow at the source. Thus, you see that without changing the load resistance or the source voltage, you can increase the current.

This completes the equivalent circuit and it appears as shown in Figure 10-1E. But it will not perform exactly like the actual line itself. An actual line is uniformly constructed so that energy going either way will encounter the same effects. This is not the case in the equivalent circuit, for if you look into the line shown at E from the left end, you see two  $R$ 's and one  $C$  connected in series, and if you look into the right end, you see two  $L$ 's and one  $G$  in series. This is not the same circuit.

To balance and simplify the circuit, it is changed as you can see in Figure 10-1F so that all the series resistance is put into the top line, giving the same electrical effect. The series resistance is then divided into halves and each half

is located as shown at F. The same is done with the inductance. The net result is that the circuit looks the same from either end.

You can further reduce the circuit to the simple impedance network as shown in Figure 10-1G. In this case all the series impedances (represented by  $Z_1$ ) are symmetrically spaced about the parallel impedance (represented by  $Z_2$ ). When the circuit is reduced to this form, it is possible to develop certain general equations in terms of  $Z_1$  and  $Z_2$ . These equations, which are discussed later, apply to any circuit regardless of the resistance, the inductance, or the capacitance in  $Z_1$  and  $Z_2$ .

Figure 10-1H shows another form into which the r-f line constants can be lumped. In this circuit the shunt constants are symmetrically placed about the series constants. This is known as a Pi section and is shown in terms of impedances in Figure 10-1I.

### The Characteristic Impedance of a Lossless Line

In an r-f line there is an inherent impedance that is constant regardless of the length or the applied voltage. In the equivalent circuit shown in Figure 10-2A, the r-f line is assumed to be without losses. Remember that inductance or capacitance will store energy, then return it to a circuit, while resistance will dissipate energy in the form of heat. Since no resistors appear in the circuit at A, no energy will be lost. A line can be constructed with extremely small losses; this approximates the ideal situation. Such a line is assumed to be infinitely long. Although this, too, is a theoretical situation, cross-country telephone lines approach the condition described.

Notice what happens when you apply a DC voltage to this circuit. At the time you close the switch, the full voltage  $E$  will appear across the line terminals A-B. This end is referred to as the *sending end*. The end to which energy travels is called the *receiving end*. The first capacitor  $C_1$  starts to charge. But its charge is not instantaneous, since the charging current is impeded by the inductance of  $L_1$ . After a short time, however,  $C_1$  will approach full charge, and its charging current will decrease.  $C_2$  will start to charge through  $L_2$ , and the resultant current through the ammeter (A) in the circuit will remain constant.



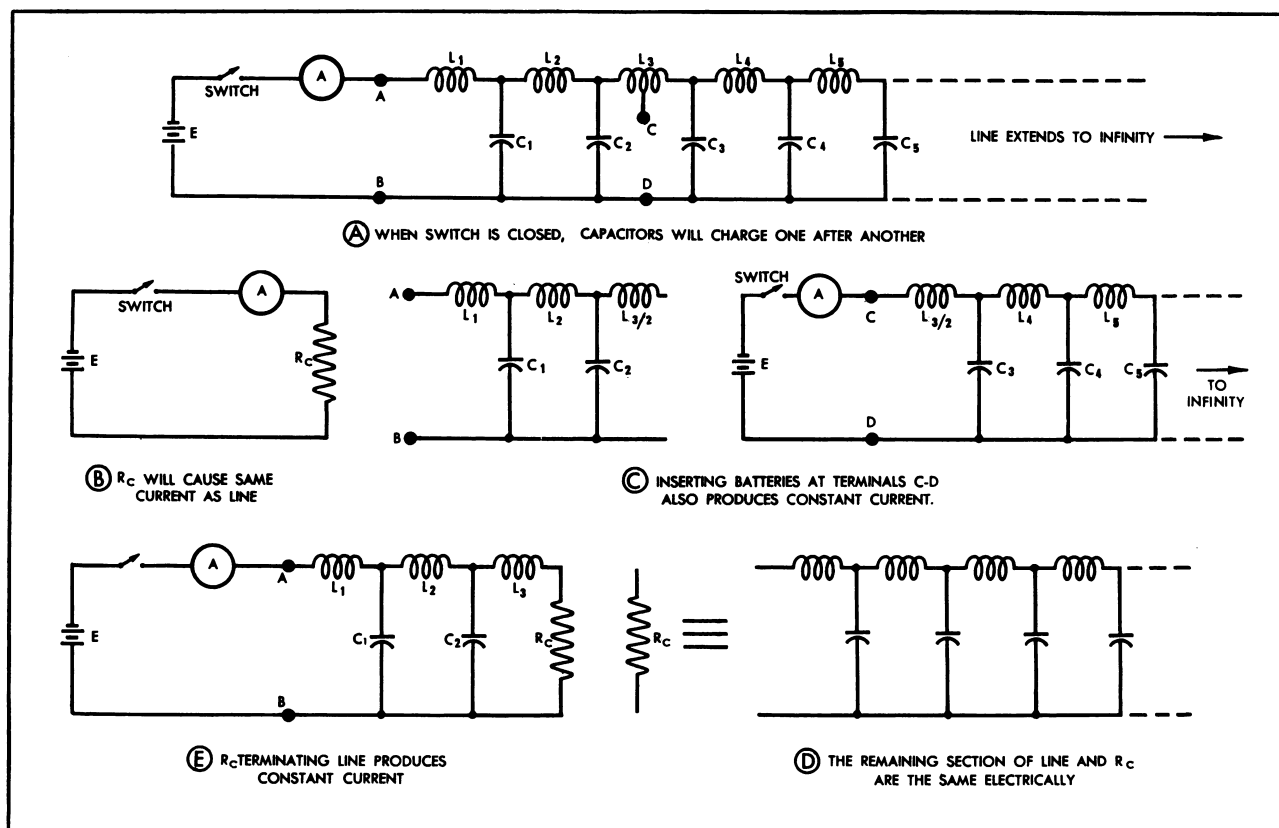


Figure 10-2. Developing Characteristic Impedance

A current will flow through the inductors to charge the next capacitor. Each capacitor down the line will add to the charging current as the current through the preceding capacitor starts to decrease. Since the line is infinitely long, there will always be more capacitors to be charged, and current will not stop flowing until they are all charged. The line has the same effect upon the battery as would a fixed resistor, with a value equal to the characteristic impedance of the line.

Another circuit that will display identical characteristics is shown in Figure 10-2B. The resistor  $R$  in this circuit will cause a certain current to flow indefinitely when voltage  $E$  is applied. (To determine the value of  $R$ , use Ohm's law formula,  $R = E/I$ .) Since the voltage  $E$  is known and a constant current  $I$  is indicated by the meter, the r-f line displays a resistance equal to the value  $E/I$ . This resistance is the characteristic resistance of the r-f line, and may be called  $R_c$ . Mathematically, it is

$$R_c = \frac{E}{I}.$$

If you broke the line at the center of  $L_3$  and connected the battery to terminals C-D as shown in Figure 10-2C, again the same current would flow, for the same constants are displayed and the receiving section is still infinitely long.

If a resistance equal to the characteristic resistance of the section removed is placed across the receiving end-terminals of the short section and the battery is again connected across terminals A-B as shown in Figure 10-2E, the same current will flow again. Therefore, whether the line is infinitely long or fixed in length and terminated in a resistance equal to its characteristic resistance, the effect at the battery end is the same. The display of constant resistance that you see at the sending end is equally important, for it is this resistance that must be matched by the generator for maximum power transfer. When AC voltages are applied to the line, the characteristics of the line may be resistive with a reactive component—either capacitive or inductive. Therefore, the characteristic of a line generally is described as the characteristic impedance, for which the symbol is  $Z_0$ .

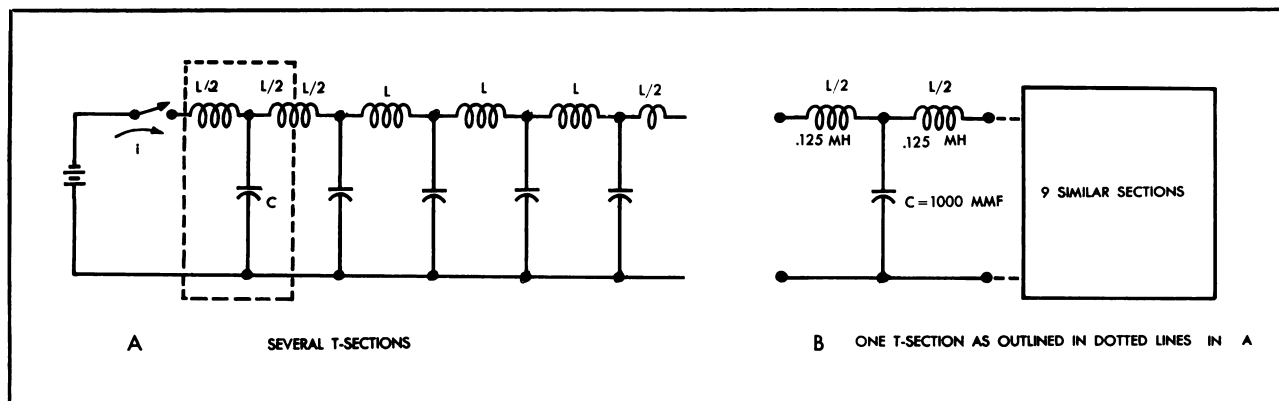


Figure 10-3. Circuit for Calculating Time for Wave to Travel Down the Line

### Voltage Changes Moving Along a Line

Previously it was stated that a voltage suddenly applied will appear at a point some distance from the sending end at some later time. This is true regardless of whether the voltage change is a jump from zero to some value, a drop from some value to zero, or for that matter, any other change in voltage. Whatever the change, the voltage will be conducted down the line at a constant rate.

If you know the inductance and capacitance of an r-f line, you can easily compute the time required by any waveform to traverse the length of the line. The circuit in Figure 10-3A is the equivalent circuit of a line reduced to T-sections and assumed to be lossless. The inductance per section is represented by  $L$ ; the capacitance per section, by  $C$ . In this connection recall a few fundamental relationships, such as  $Q = IT$ , which means that the quantity of electricity in coulombs is equal to the current multiplied by the time the current flows, and  $Q = CE$ , which means that the quantity of electricity in a capacitor is equal to the capacitance times the voltage across it. In the case of the transmission line shown at A, the  $Q$  of electricity leaving the battery is also shown by the equation,

$$Q = IT \quad (1)$$

where  $I$  is the current and  $T$  the time the current flowed.

The electricity that leaves the battery goes into the line where it is deposited in the capacitors as a charge ( $Q$ ). This charge is shown by the equation,

$$Q = CE \quad (2)$$

where  $C$  is the capacitance per section and  $E$  the voltage applied.

The quantity of electricity that left the battery is equal to the quantity in the line; that is,  $Q = Q$ . Therefore, by equating (1) and (2),

$$IT = CE \quad (3)$$

As each capacitor charges to  $CE$ , the voltage at the end of the inductor must change by  $E$  with respect to the other end. This occurs because the first capacitor charges to  $E$  voltage, while the second capacitor is still at zero voltage. This makes the voltage  $E$  appear across the inductor. In an inductor these circuit components are related mathematically as,

$$E = L \frac{\Delta I}{\Delta T} \quad (4)$$

The voltage across the inductor is directly proportional to the inductance and the change in current but increases as the time becomes shorter. This is expressed mathematically by the equation,

$$E\Delta T = L\Delta I \quad (5)$$

which is the result of multiplying the preceding equation by  $\Delta T$ .

Since current and time start from zero, the change in time ( $\Delta T$ ) and the change in current ( $\Delta I$ ) are equal to the final time  $T$  and final current  $I$ . For this case, equation (5) becomes,

$$ET = LI \quad (6)$$

If voltage  $E$  is applied for time  $T$  across inductor  $L$ , the final current  $I$  will flow. The following equations contain all three terms ( $T$ ,  $L$ , and  $C$ ).

$$IT = CE \quad (3)$$

$$ET = LI \quad (6)$$

Another way of using them is to get an equation for  $T$  in terms of  $L$  and  $C$ . This is done in the following manner: Multiply the left and right members of each equation as follows:

$$(IT)(ET) = (CE)(LI) \quad (7)$$

$$EIT^2 = LCEI$$

$$\text{Cancelling the EI terms, } T^2 = LC \quad (8)$$

$$\text{and} \quad T = \sqrt{LC} \quad (9)$$

Equation (9) is the equation for the time required for a voltage change to travel along a section composed of  $L$  and  $C$ .

Since the characteristic impedance is resistive and equal to  $R = E/I$ , it is equal to the voltage-to-current ratio. Since  $E$  and  $I$  appear in the preceding equations, these equations can be solved for the  $E/I$  ratio. This is done as follows:

Divide equation (6) by equation (3):

$$\frac{ET}{IT} = \frac{LI}{CE} \quad (10)$$

Multiply both sides by  $\frac{E}{I}$ :

$$\frac{E^2 T}{I^2 T} = \frac{L}{C} \quad (11)$$

$$\text{Cancel the } T\text{'s: } \frac{E^2}{I^2} = \frac{L}{C} \quad (12)$$

$$\text{Take square root: } \frac{E}{I} = \sqrt{\frac{L}{C}} \quad (13)$$

$$\text{Since } E/I \text{ is the } Z_o \text{ of the line, } Z_o = \sqrt{\frac{L}{C}} \quad (14)$$

### Example

#### Problem

Assume that the line illustrated in Figure 10-3A is 1000 feet long and that a 100-foot section is measured and found to have an inductance of  $\frac{1}{4}$  millihenry and a capacitance of 1000 mmf as shown at B. Find the characteristic impedance of the line and the time required for a voltage to travel the length of one section.

#### Solution

$$Z_o = \sqrt{\frac{L}{C}} = \sqrt{\frac{0.25 \times 10^{-3}}{10^{-9}}} = \sqrt{0.25 \times 10^6} = 0.5 \times 10^3$$

$$Z_o = 500 \text{ ohms}$$

The time required for a voltage to travel the length of one section is equal to,

$$T = \sqrt{LC} = \sqrt{0.25 \times 10^{-3} \times 10^{-9}} = \sqrt{0.25 \times 10^{-12}} \\ = 0.5 \times 10^{-6} \text{ sec.}$$

$$T = 0.5 \text{ microsecond}$$

The time required for voltage to travel over ten sections (1000 feet) is equal to  $0.5 \times 10 = 5$  microseconds.

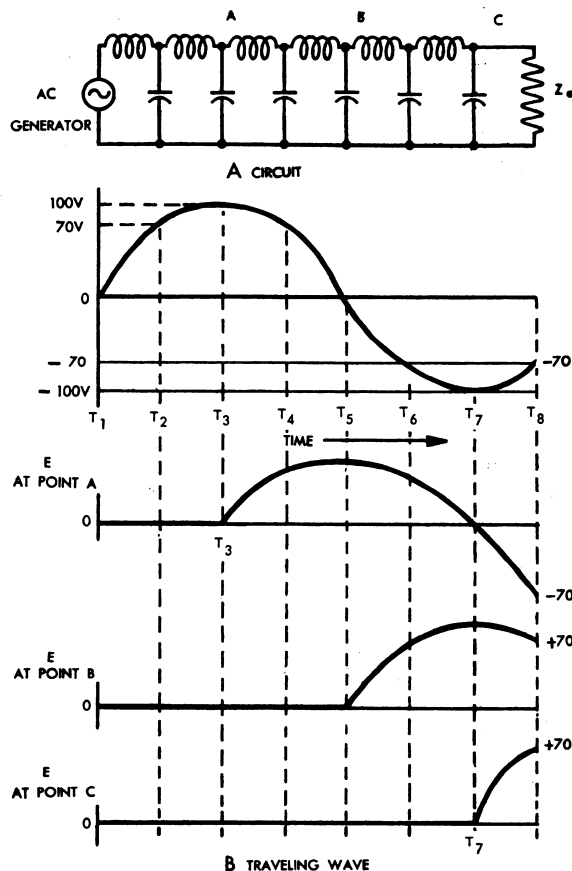


Figure 10-4. Sinusoidal Voltage Travels Down Line

This example was based on a DC change. An r-f line displays similar characteristics when an AC voltage is applied to its sending end terminals. Assume that an AC voltage is applied to the line simulated by the circuit shown in Figure 10-4A. The generator voltage starts from zero, and produces the voltage indicated. As soon as a small voltage change is produced, it starts its journey down the line, and the generator in the meantime continues to produce new voltages along a sine curve. At time  $T_1$ , the generator voltage is 70 volts. The voltages still move along the line. Finally, at time  $T_1$ , the first small change arrives at point A, and the voltage at that point starts increasing. At time  $T_2$ , the same voltage arrives at point B on the line, and finally at time  $T_7$ , the first small change arrives at the receiving end of the line. Meanwhile all the changes in the sine wave pass each point in turn. The time required for the changes to travel the length of the line is the same time required for a DC voltage to travel the distance.

At time  $T_7$ , the voltages at the various points on the line are as follows:

|               |       |
|---------------|-------|
| At generator: | -100v |
| At Point A:   | 0v    |
| At Point B:   | +100v |
| At Point C:   | 0v    |

If you plot these voltages along the length of the line, the resulting curve is like the curve shown in Figure

10-5A. Note that such a curve of instantaneous voltages resembles a sine wave.

At time  $T_1$  in Figure 10-4, the voltages line up in the following manner:

|               |                            |
|---------------|----------------------------|
| At generator: | Drop from $-100$ to $-70v$ |
| At Point A:   | Rise from zero to $-70v$   |
| At Point B:   | Drop from $100$ to $70v$   |
| At Point C:   | Rise from zero to $70v$    |

A plot of those voltages produces the solid curve shown in Figure 10-5B. For reference, the curve from time  $T_1$  is drawn as a dotted line. The solid curve has exactly the same shape as the dotted curve, but it has moved bodily down the line the distance  $X$ . Another plot at  $T_2$  would show a new curve similar to the one at  $T_1$  but moved to the right by the distance  $X$ .

### Summary

By analyzing the points just discussed, you will find that the actions associated with voltage changes along an r-f line are as follows:

1. All parts of the sine wave produced by the generator travel down the line in the order produced.
2. At any point you will obtain a sine wave if you plot all the instantaneous voltages passing the point during eight of the time intervals previously shown. The oscilloscope will plot all values instantaneously.
3. The oscilloscope pictures (instantaneous voltages) will be the same in all cases except that a phase difference will exist in each picture taken down the line. The phase will change continually with respect to the generator until over a certain length of line the change will be  $360^\circ$ .
4. Since all parts of a sine wave pass every point along the line, a plot of the readings of an AC meter (which reads the magnitude of the voltage over a period of time) will show that the voltage is constant at all points as shown in Figure 10-5C.
5. Since the line is terminated with  $Z_0$ , the energy arriving at the end is absorbed by the resistance.

The distance traveled by the start of one cycle, during the time required to generate the entire cycle, is one wavelength. In Figure 10-6 the voltage shown by the solid line arrives at the receiving end at the same time a similar voltage is generated in the next cycle. This line is exactly one electrical wavelength long and contains a full wave of voltage at one time.

The electrical length of a line is a function of the frequency of applied voltage and the velocity

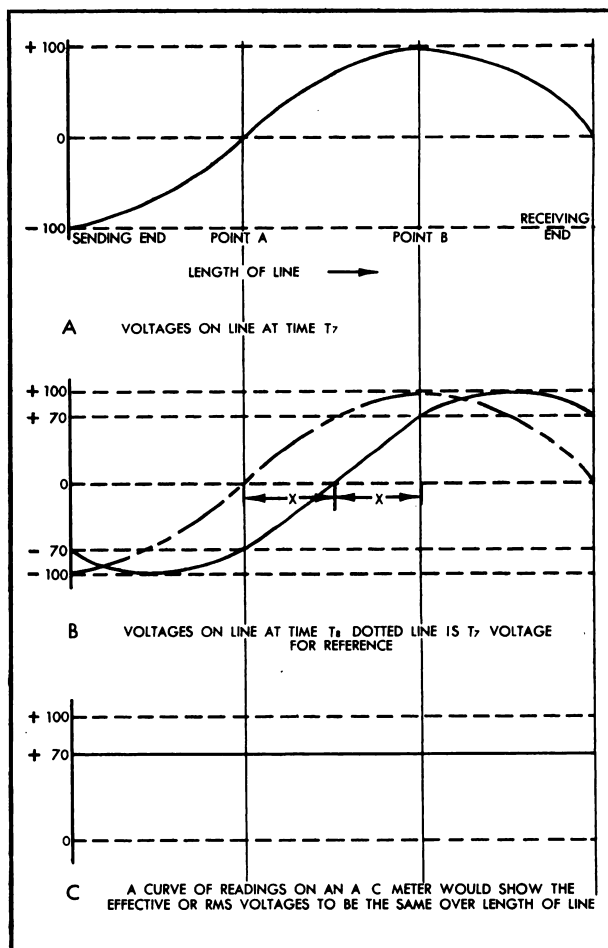


Figure 10-5. Instantaneous Voltages are Different Along Length of Line Voltages for Moving Sine Wave

of energy travel. The time required for travel per section is equal to,

$$T = \sqrt{LC} \quad (9)$$

Velocity is the distance per unit of time and distance  $D$  is the length of one section, thus,

$$v_g = \frac{D}{T} \quad (15)$$

The subscript of  $g$  indicates velocity of waves on a line as distinguished from velocity of waves in space. In terms of  $L$  and  $C$ , equation (15) may be expressed,

$$v_g = \frac{D}{\sqrt{LC}} \quad (16)$$

If, for example,  $\sqrt{LC}$ , or  $T$ , is 0.5 microsecond, and the section is 100 feet long, the velocity of

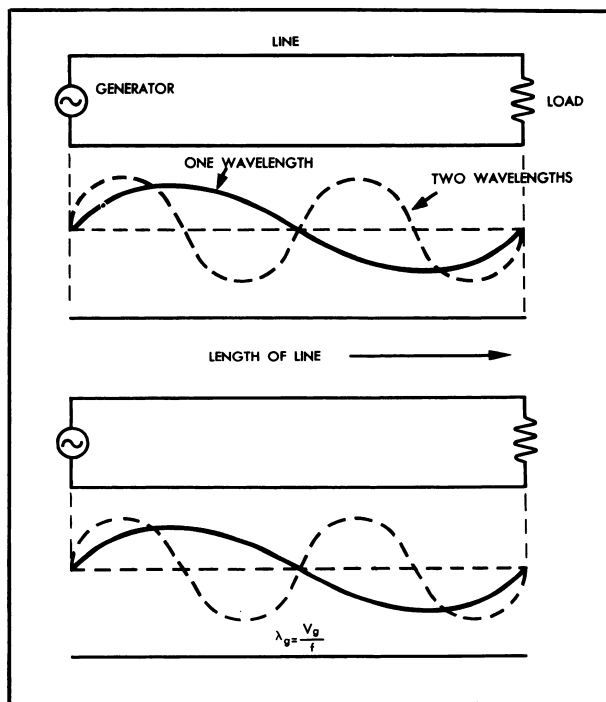


Figure 10-6. Meaning of Wavelength

the wavefront,  $V_g$ , is found as follows:

$$V_g = \frac{D}{\sqrt{LC}} = \frac{100}{0.5 \times 10^{-6}} = \frac{100 \times 10^6}{0.5}$$

$V_g = 200 \times 10^6$  or 200 million feet per second or approximately 39,000 miles per second.

The wavelength (of the group) of voltages on the line is found by the equation,

$$\lambda_g = \frac{V_g}{f} \quad (17)$$

where  $f$  is the frequency in cycles per second.  $\lambda_g$  is wavelength in feet if  $V_g$  is velocity in feet per second.

#### Example

##### Problem

Assuming that the generator frequency in the preceding example is 200 kc, find  $\lambda_g$ .

##### Solution

$$\lambda_g = \frac{V_g}{f} = \frac{200 \times 10^6}{200 \times 10^3} = \frac{10^6}{10^3} = 10^3$$

$$\lambda_g = 1000 \text{ feet}$$

One wavelength will occupy 1000 feet of the line. Notice that if you double the frequency, the wavelength is only one-half of the former value. A 1000-foot line will have two full cycles of 400-kc voltage on it at any one time. This is shown by dotted line in Figure 10-6.

Little has been said thus far about the current changes in the r-f lines. The r-f lines described were all assumed to display a resistance to the current, but no reactance to it, and the termination to have the characteristic impedance of the line. Under these conditions, the current is always directly proportional to the voltage, and whenever a voltage appears across a point on the line, a current equal to  $E/Z_0$  will immediately flow at that point as shown in Figure 10-7. As a DC voltage change moves along the line, a current change moves with it. As an AC voltage moves along the line, a current exactly in phase with it is present at all points.

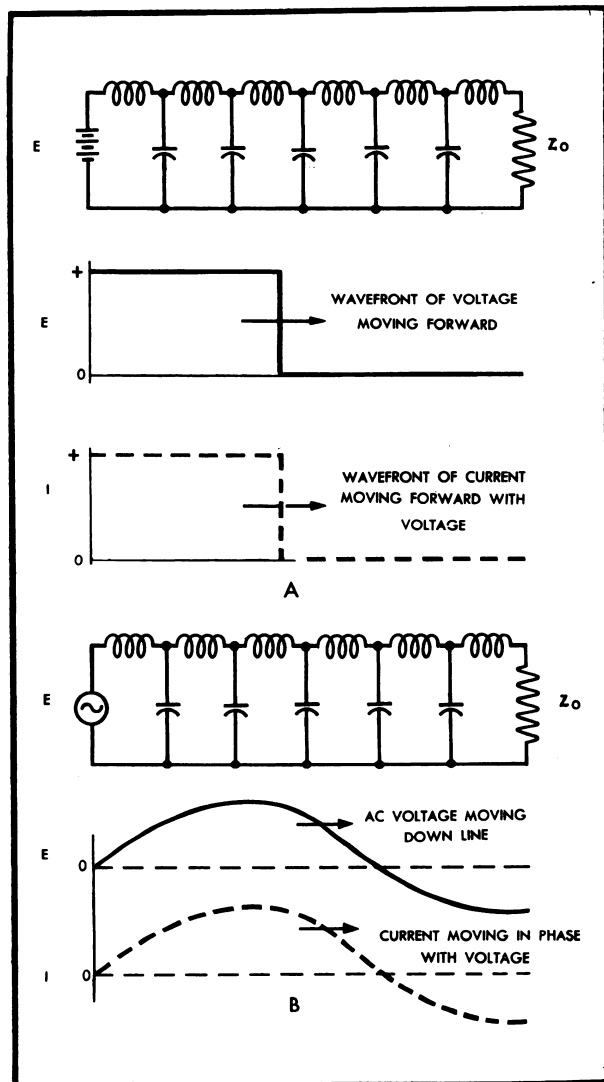


Figure 10-7. Current Changes Move in Step with Voltage

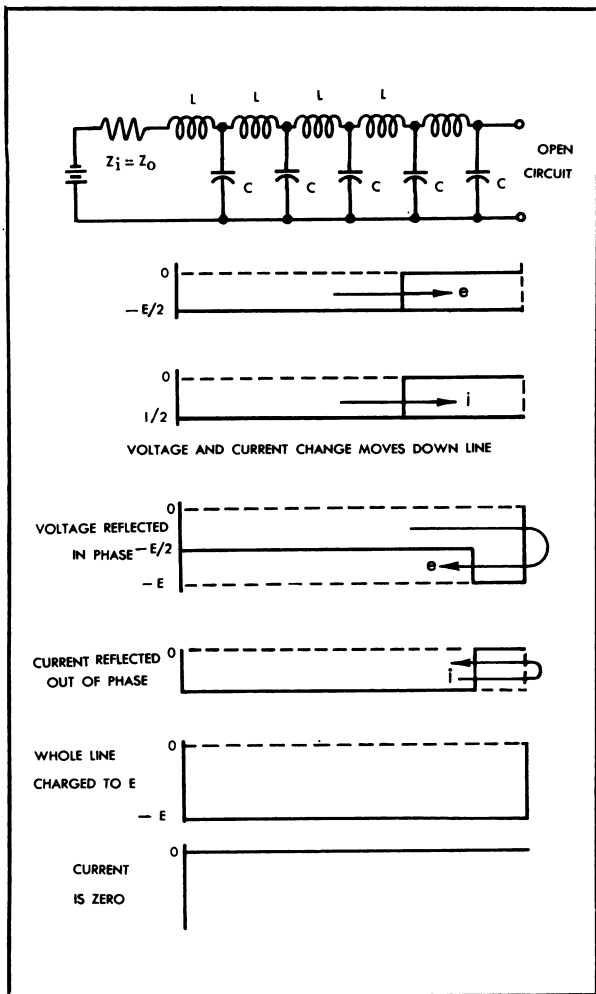


Figure 10-8. Reflection from Open-Ended Line

## REFLECTION

It was assumed in the analysis of r-f lines that the terminating resistor absorbed all the energy. However, r-f lines are not always terminated in this manner. In some cases you will find the receiving end open-circuited or short-circuited.

Take the case of the line which has an open circuit at the receiving end and which has DC voltage applied at the sending end. As the switch in the equivalent line shown in Figure 10-8 is closed at the sending end, the negative voltage moves down the line, charging each capacitor through the preceding inductor. When the last capacitor is charged, there is no longer any voltage change. The capacitor at each end of the inductor is charged to  $\frac{1}{2}E$  (since  $Z_i = Z_0$ , one-half the voltage drop is across  $Z_i$  and one-half

across  $Z_0$ ). As there is no voltage across the inductor to maintain the magnetic field, the field collapses and causes current to flow into the last capacitor. Furthermore, as the direction of current has not changed, the capacitor charges in the same direction.

Since the energy in the magnetic field described equals the energy in the capacitors, the transfer to the capacitor is just enough to double the voltage across the last capacitor. The current in the inductor drops to zero. Since there is no place for current to flow from the next to the last inductor, its field collapses, doubling the voltage across the next to the last capacitor. This change of voltage (moving backward) is the same as though the voltage on arriving at the end found no place to go and started back in the same polarity. Such action is called *reflection*. A reflection of voltage at an open always occurs in the same polarity. The voltage change moves back to the source, charging each capacitor in turn until the last capacitor is charged to the source voltage and the action stops.

In the meantime the field collapsed around the coil and the current there became zero. As each capacitor is charged, the current in each inductor drops to zero, effectively reflecting the changed current in opposite polarity. The opposite polarity cancels the current at each point and it drops to zero at that point. When the last capacitor is charged, the current from the source stops flowing.

Notice that the line current flowed through the resistor called  $Z_i$ . This is the internal resistance of the source. For maximum power transfer to the line,  $Z_0$  of the line must equal  $Z_i$ . Because of this resistance, the voltage change delivered to the line is one-half of the source voltage. On the return trip, the additional half charges each capacitor to the source voltage.

The principal facts in the reflection of voltages in open-end lines are:

- Voltage is reflected in phase and without change in amplitude or shape from an *open* end.
- Current is reflected in opposite phase and without change of amplitude or shape from an *open* end.

## Reflection from Short Circuit

The effect of a short-circuited line on voltage change is different from that of an open-circuited line. In a short-circuited line the voltage change

from zero to  $E/2$  arrives at the last inductor as usual, but in this case the current in the inductor has no capacitor to charge. As you can see in Figure 10-9C, the current in the final inductor causes the indicated polarity to exist. When the field collapses, the inductor becomes a source and the polarity reverses. In the instantaneous equivalent circuit at D showing this action, the inductor acts as a battery and forces current through the capacitor in the opposite direction, causing it to discharge. Since the energy in the magnetic field is the same as that in the capacitor, the capacitor discharges exactly to zero. The second inductor current consequently has no voltage to maintain it; therefore, it will discharge the inductor which is next to the last one.

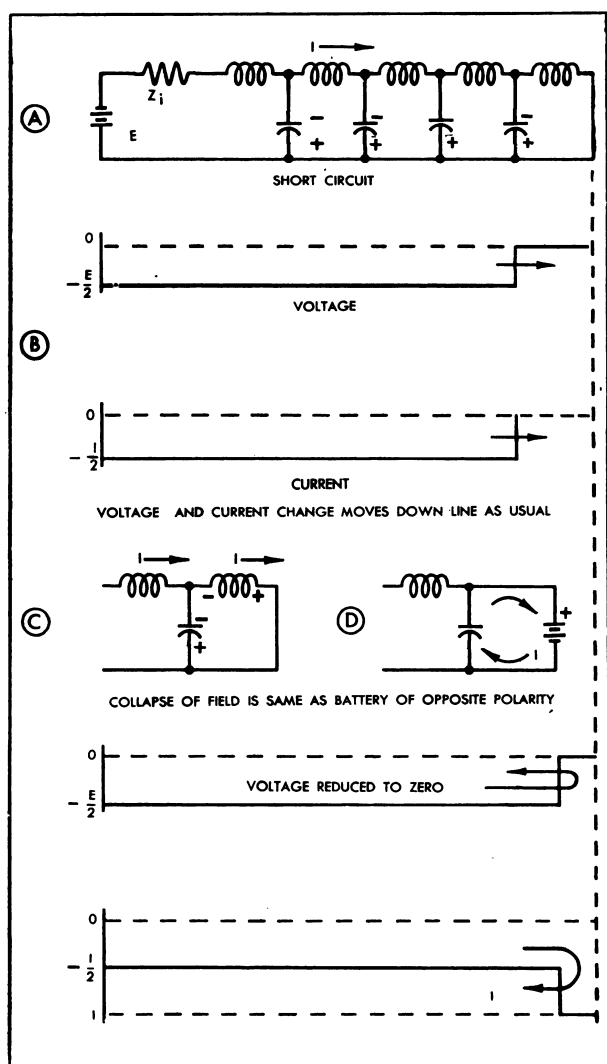


Figure 10-9. Reflection from Short-Circuited Line

The reduction of voltage to zero by the inductor voltage becomes effectively a new voltage that appears at the end of the line. It is equal to  $E/2$ , but has the opposite polarity. The collapsing field adds a current which, in turn, adds to the current created by the source. In effect, the current change appears at the end of the line in the same direction as the original current.

An r-f line with a short circuit across it is characterized by:

- Voltage changes that are reflected in opposite polarity but with the same amplitude.
- Current changes that are reflected in the same polarity and with the same amplitude.

#### Reflection of AC on an Open-Ended R-F Line

In most cases where r-f lines are used, the voltages that are applied to the sending line are AC voltages. When AC voltages are applied,

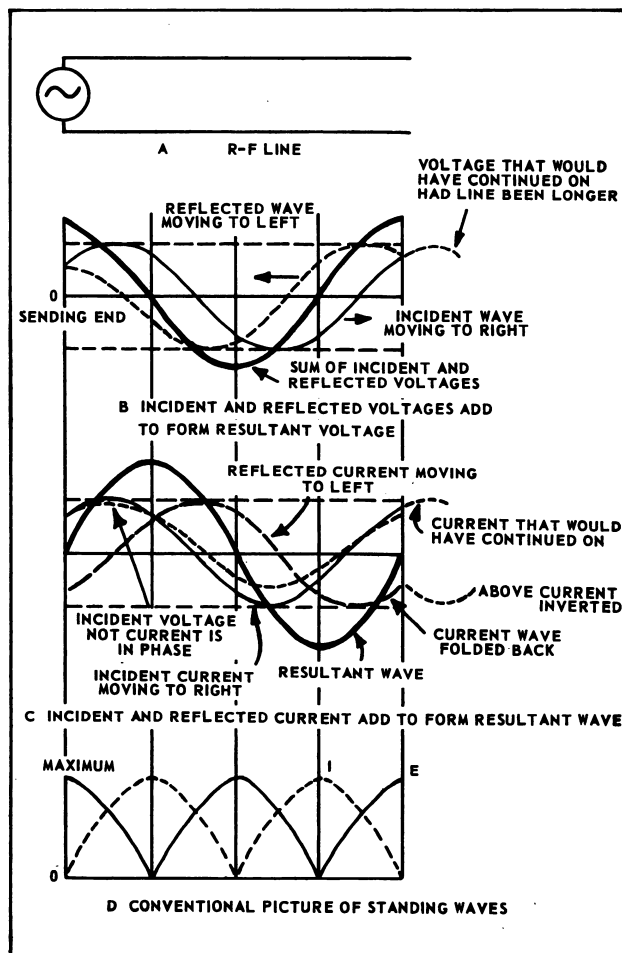


Figure 10-10. Formation of Standing Waves

the action at the receiving end of the line is exactly the same as for DC. In the open-ended line, shown in Figure 10-10A the generated AC voltage is distributed along the line (as shown at B) in such a way that as it arrives at the end, it is reflected with the same polarity and amplitude. When AC is used, this reflection is in phase. The reflected voltage wave travels back along the line until it reaches the generator. If the generator impedance is the same as the line impedance, energy arriving at the generator is absorbed. There are now two voltages on the line. However, most indicating instruments are unable to indicate these voltages separately. Instead, they read their vector sum.

An oscilloscope is usually used to study the instantaneous voltages on r-f lines. It is used because meters read average values over a period of time. However, you may first add the two voltages algebraically on paper as in Figure 10-10B to see how they will appear on the oscilloscope.

Since there are two waves of voltage moving on the line, it is desirable to distinguish between the two. This is done by giving them names. The voltages moving toward the receiving end are called *incident voltages*, and the whole wave-shape is called the *incident wave*. The wave moving back to the sending end after reflection is called the *reflected wave*.

Another step in investigating the open-ended r-f line is to see how the current waves act. Note that the incident current wave is the solid line in Figure 10-10C. The voltage at the moment is represented by the dotted line. The current is in phase with the voltage while traveling toward the receiving end. At the end of the line the current is reflected in opposite polarity; that is, it is shifted  $180^\circ$  in phase, but its amplitude remains the same. The reflected wave of current is shown by dashed lines in Figure 10-10C.

Since a current indicating device would read the sum of the two currents, you must add them to determine what the current indicator will show. The heavy-line curve represents the sum of the two currents. Notice that current is zero at the end of the line. This is reasonable since there can be no current flow through an open circuit. The voltage curve shows that the voltage is maximum at the end of the line, a condition that can easily occur across an open circuit.

Another factor to keep in mind is that when you use an AC meter to determine the current or voltage on an r-f line, it will indicate only the magnitude of the current, not the polarity. If you plot all the readings along the length of the line, you will obtain a curve like the one shown in Figure 10-10D. Notice that all the readings are positive due to current rectification in the meter circuit. The curves in Figure 10-10D represent the conventional method of showing current and voltage on an r-f line.

The voltages and currents shown at B and C in Figure 10-10 are those at a single instant. During the generation of a complete cycle by the generator, there are a large number of these pictures generated. The instantaneous incident and reflected waveshapes are shown in Figure 10-11.

Drawings 1 through 8 in Figure 10-11 show reflection in phase. This is the case with voltage when the line is open-circuited at the end. If you check each picture individually, you will see that the heavy curve is the algebraic sum of the other two curves. In going from picture to picture, you can see that one curve moves to the right and the other moves to the left. The curve moving to the right is the incident wave of voltage; the one moving to the left is the reflected wave of voltage. The composite picture at A in Figure 10-12 shows all resultant curves over a complete cycle. Notice that the voltage varies between zero and maximum in both directions at the center and at both ends as well. But the voltage is always zero at intermediate points, in this case, one-fourth of the distance from each end. The resultant waveshape, which is not traveling like the other waves, is referred to as a *standing wave* of voltage.

The voltage at the center and the ends varies at a sinusoidal rate between the limits shown and is always at zero at the one-quarter and three-quarter points. A continuous series of pictures such as these are difficult to see with conventional test equipment, for this equipment reads the effective or average voltage over a number of cycles. The curve of amplitudes over the length of line for several cycles is shown in Figure 10-12B. A meter would read zero at the points shown and show a maximum voltage at the center no matter how many cycles went by.

The current waveshapes on the open-ended line are shown in the right-hand column in



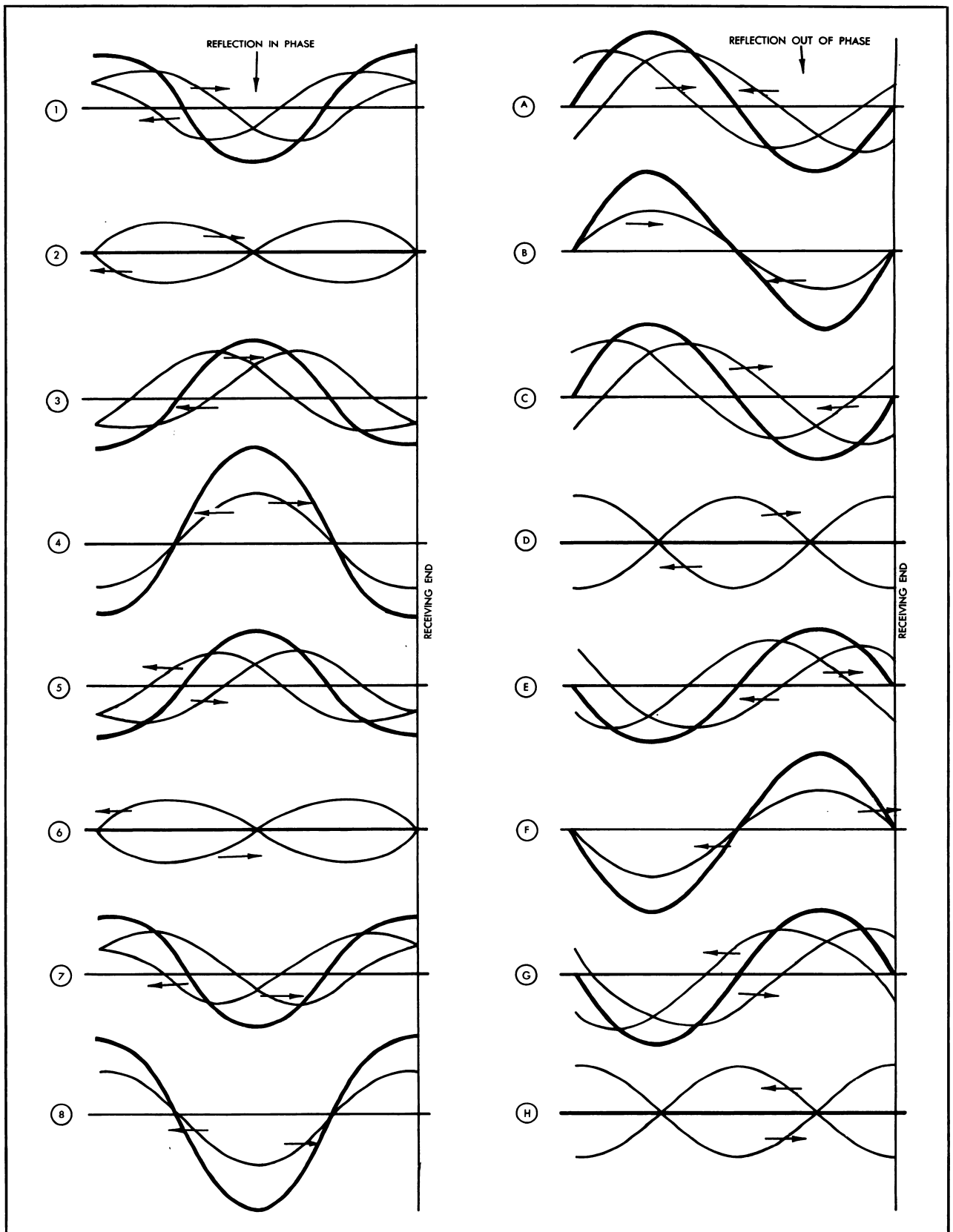


Figure 10-11. Instantaneous Values of Incident and Reflected Waves

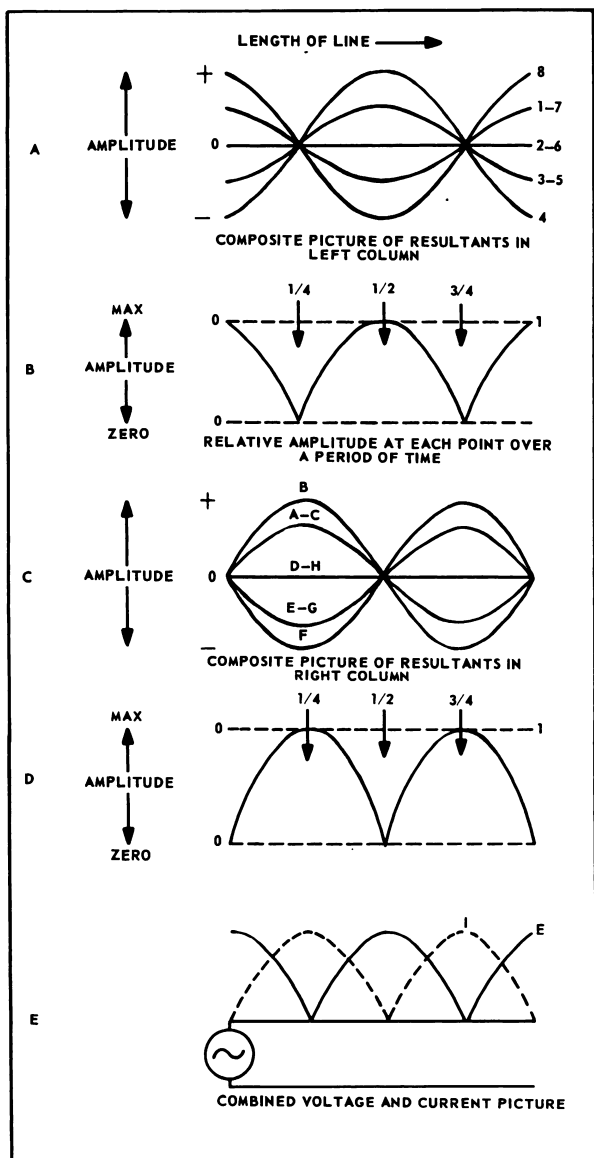


Figure 10-12. Composite Results of Instantaneous Waves

Figure 10-11. Since the reflection of current is out of phase at an open end, the resultant wave-shapes differ from those on the left side. The two out-of-phase components always cancel one another at the end, and thus the resultant is always zero at that point. (See the composite picture of resultants in Figure 10-12C.) Again the voltage is zero at certain points at all times, but maximum at other points. Therefore, it is a standing wave. As shown at D, meters would show the amplitude along the length. In this case, it is zero at the end and center, but maximum at the one-quarter and three-quarter points.

The entire picture of the open-circuited line conditions is usually as shown in Figure 10-12E. The standing waves of voltage and current appear together. Observe that one is maximum when the other is minimum. The current and voltage standing waves are one-quarter cycle, or  $90^\circ$ , out of phase with one another.

#### Reflection of AC on a Short-Circuited R-F Line

We have stated before that when an r-f line is terminated in a short circuit, reflection is complete but the effect on voltage and current differs from the effect in an open-circuited line. Voltage is reflected in opposite phase, while current is reflected in phase. Again refer to the series of pictures shown in Figure 10-11. The left column represents current since it shows reflection in phase. The right column of pictures shows the voltage changes on the shorted line. The final effect of this is shown in Figure 10-13. It shows the voltage as a solid line in which the voltage is zero at the end and center, and maximum at the one-quarter and three-quarters points.

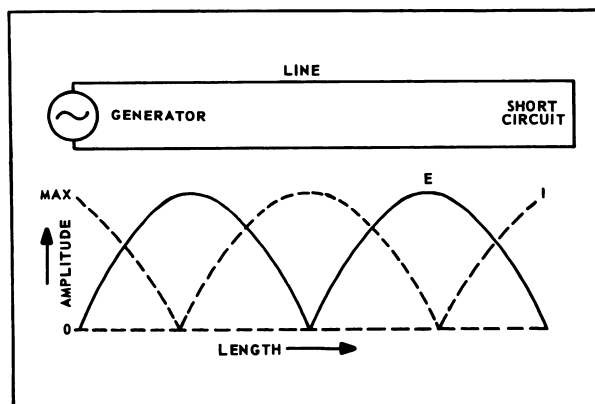


Figure 10-13. Standing Waves on Shorted Line

#### COMPARISON OF STANDING WAVES WITH VARIOUS TYPES OF TERMINATIONS

There are many different terminations for r-f lines. Each type of termination has a characteristic effect on the standing waves on the line. From the nature of the standing waves, you can determine the type of termination which produced the waves.

## Termination in $Z_0$

Termination in  $Z_0$  (characteristic impedance) will cause a constant reading on an AC meter when it is moved along the length of the line. As illustrated in Figure 10-14A, the curve for such a condition, provided there are no losses

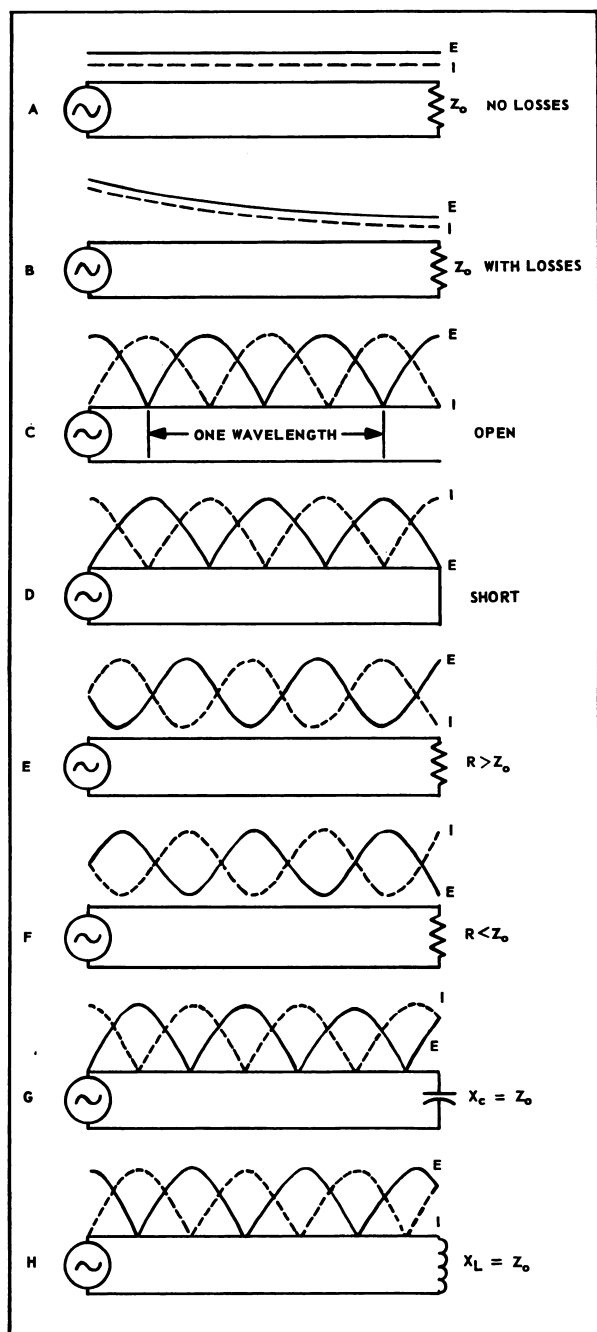


Figure 10-14. Effects of Various Terminations on Standing Waves

in the line, will be a straight line. If there are losses in the line, the amplitude of the voltage and current will diminish as they move down the line. The losses are due to DC resistance in the line itself. As the DC losses in radar equipment are often small enough to be disregarded, it is practical to assume no losses.

## Termination in Open Circuit

In an open-circuited r-f line (Figure 10-14C), the voltage is maximum at the end, but the current is minimum. The distance between the two zero points is one-half wavelength, and the distance between alternate zero points is one wavelength. It can be said that the voltage is zero at a distance of one-quarter wave from the end of the line. (This is true at any frequency.) A voltage peak occurs at the end of the line, and also one-half wavelength from the end.

## Termination in Short Circuit

On the line terminated in a short circuit (Figure 10-14D) the voltage is zero at the end and maximum one-quarter wavelength from the end.

## Termination in a Resistance not Equal to the Characteristic Impedance

Whenever the termination is anything but  $Z_0$ , there are reflections. For example, if the terminating element contains resistance, it will absorb some energy, but if the resistive element does not equal the  $Z_0$  of the line, some (not all) of the energy will be reflected. The amount of voltage reflected may be found by the equation,

$$E_R = E_i \frac{R_L - Z_0}{R_L + Z_0} \quad (18)$$

Where  $E_R$  is reflected voltage,  $E_i$  is incident voltage,  $R_L$  is the terminating resistor, and  $Z_0$  the characteristic impedance of the line.

If you try different values in the preceding equation, you will find that the reflected voltage is always less than the incident voltage except when  $R_L$  is zero or is infinitely high. ( $E_R$  and  $E_i$  are equal.) A smaller value of reflected voltage causes a standing wave of smaller variation.

When  $R_L$  is larger than  $Z_0$ , the end of the line is somewhat like an open circuit. Thus, the standing wave of voltage will be greatest at the end of the line, while the current will be at a minimum (but not zero). (See Figure 10-14E.)

When  $R_L$  is less than  $Z_0$  as shown in Figure 10-14F, the line is terminated approximately as a short circuit. In this case the voltage is minimum at the end of the line (this fact may be verified by the preceding equation), and the value of  $E_R$  becomes negative, subtracting from  $E_i$  at the end.

The variation of standing waves will show how near the r-f line is to being terminated in  $Z_0$ . A wide variation in voltage along the length means a termination far from  $Z_0$ . A small variation means termination near  $Z_0$ . Thus, the ratio of the maximum standing wave voltage to the minimum is a measure of the perfection of the termination of a line. It is called *standing-wave ratio* and is always expressed in whole numbers. For example, a ratio of 1:1 describes a line terminated in its characteristic impedance ( $Z_0$ ).

#### Terminating the Line in Capacitance

When a line is terminated in capacitance, the capacitor does not absorb energy, but returns it all to the circuit. This means 100% reflection. The current and voltage relationships are somewhat more involved than in previous types of termination. For purposes of understanding, assume that the capacitive reactance is equal to the  $Z_0$  of the line. The current and voltage will be in phase upon arriving at the end, but in flowing through the capacitor and  $Z_0$  connected in series with it, they are shifted in phase relationship.

Thus, current and voltage arrive in phase and leave out of phase, due to the shift introduced by the capacitor. This results in the standing wave configuration shown in Figure 10-14G. The standing wave of voltage is minimum at a distance of exactly one-eighth wavelength from the end. If the capacitive reactance is greater than  $Z_0$  (smaller capacitance), the termination looks more like an open circuit; the voltage minimum moves away from the end. If the capacitive reactance is smaller than  $Z_0$ , the minimum moves toward the end.

#### Termination in Inductance

When the line is terminated in inductance, both the current and voltage are shifted in phase upon arriving at the end of the line. When  $X_L$  is equal to  $Z_0$ , the resulting standing waves are as shown in Figure 10-14H. The current mini-

mum is located one-eighth of a wavelength from the end of the line. When the inductive reactance is increased, the open circuit situation is approached and all standing waves appear closer to the end. When the inductive reactance is decreased, the standing waves move away from the end of the line.

#### Summary

A line does not have to be of any particular length to produce standing waves, but it cannot be an infinite line. The only requirement for the production of standing waves is that reflection occur; the line must not be terminated in its characteristic impedance. Reflection occurs in lines terminated in opens, shorts, capacitances and inductances, because no energy is absorbed by the termination. Because energy exists across zero resistance which cannot absorb it, the energy has no place to go except back toward the source.

The voltage and current relations for open-end and shorted lines are opposite to each other, as shown in Figures 10-14C and 10-14D. The points of maximum and minimum voltage and current are determined from the output end of the line, because reflection always begins at that end.

#### SENDING-END IMPEDANCE OF LINES OF VARIOUS LENGTHS AND TERMINATIONS

When a line is terminated in something other than  $Z_0$ , the impedance at the generator (sending end) varies widely from the characteristic impedance. An example of the standing waves on a line of random length terminated in a short circuit is shown in Figure 10-16A. The voltage is low and the current is high at the end of the line. Ohm's law shows that the impedance at this point must be low. The voltage and current distribution on the line varies. Where the voltage is high, the current is low; where the current is high, the voltage is low.

The impedance along the line will vary, tending to be large where the voltage is high and small where the voltage is low. At a point a quarter-wavelength from the end of the line, the voltage and current have shifted a quarter-cycle. The voltage here is high and the current is low. At this point the impedance is high. This will also be true of all points that are an odd multiple

of quarter-waves from the receiving end of a shorted line.

At a point a half-wavelength from the end of the line both the current and the voltage have shifted one-half cycle. The voltage is low and the current is high at all distances from the receiving end that are even multiples of a quarter-wavelength (multiples of a half-wavelength) from the receiving end. At these points the impedance is low.

If the transmission line were cut at a point a quarter-wave from the receiving end of the line, a generator inserted at this point would see a high impedance. Similarly, if the generator were inserted at the half-wave point, it would see a low impedance. The impedance along a line at any point is the impedance that would be obtained if the transmission line were cut at that point, and the impedance looking toward the receiving end were measured with the proper equipment.

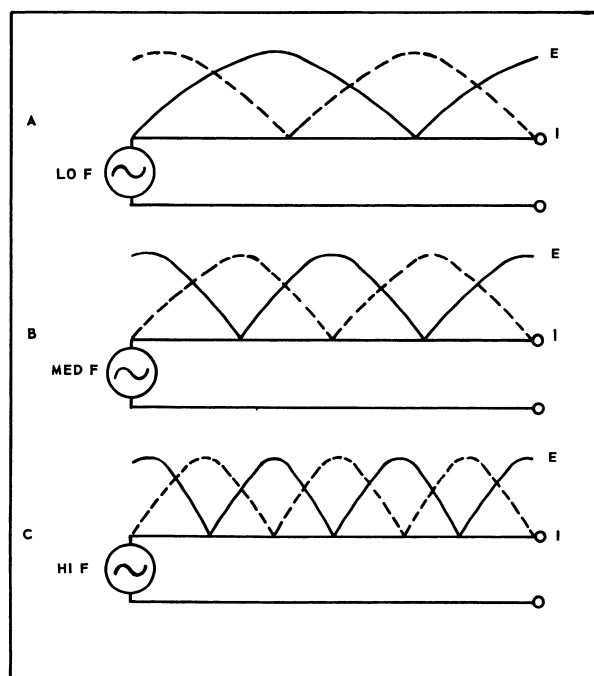


Figure 10-15. Effect of Frequency on Electrical Length of Line

What happens if frequency of a generator working into a transmission line increases or decreases? Figure 10-15B illustrates a generator feeding a full-wave line. If the frequency of the generator is decreased, the lower frequency pro-

duced has a longer wavelength. The same physical length of transmission line is now not long enough to have a complete cycle on it. Figure 10-15A shows the same transmission line as used in B. Notice that now, at the lower frequency, the line is approximately three-quarters of a wavelength.

Similarly, the electrical length of the line will change if the frequency of the generator increases. Figure 10-15C again shows the same piece of transmission line with a generator of higher frequency feeding it. A higher frequency will have a shorter wavelength. Therefore, more cycles of the standing waves will appear on the line. The line appears to the high-frequency generator as approximately one and a half wavelengths.

### Power Factor of Transmission Lines

The power factor of the impedance of a transmission line also varies. Power factor, as you know, is the cosine of the phase angle between the voltage and the current. Where there is no reflected wave, the power factor of the impedance at the point concerned and the power factor of the characteristic impedance of the line are equal. We have shown that there is no reflected wave when the line is terminated in its characteristic impedance. At the high frequencies used in radio and radar, the power factor of such a line is resistive, that is, equal to unity. A generator inserted at any point in a "flat" line (a line terminated in its  $Z_0$ ) will always see the characteristic impedance of the line.

When the line is not terminated in its characteristic impedance, there is a reflected wave. The power factor is resistive only at the points of minimum and maximum voltage. On the lines shown in Figures 10-16A and 10-16B this would be at the quarter-, half-, and three-quarter-wave points. Between the points of unity power factor, the power factor is reactive. The line in Figure 10-16A, which is terminated in a short, is inductive from the receiving end to the first voltage maximum. At this point, as we have explained, the power factor is resistive.

Immediately beyond this resistive point, the power factor along the line is capacitive as far as the voltage minimum point, where the power factor is again suddenly resistive. The power factor is alternately inductive and capacitive for

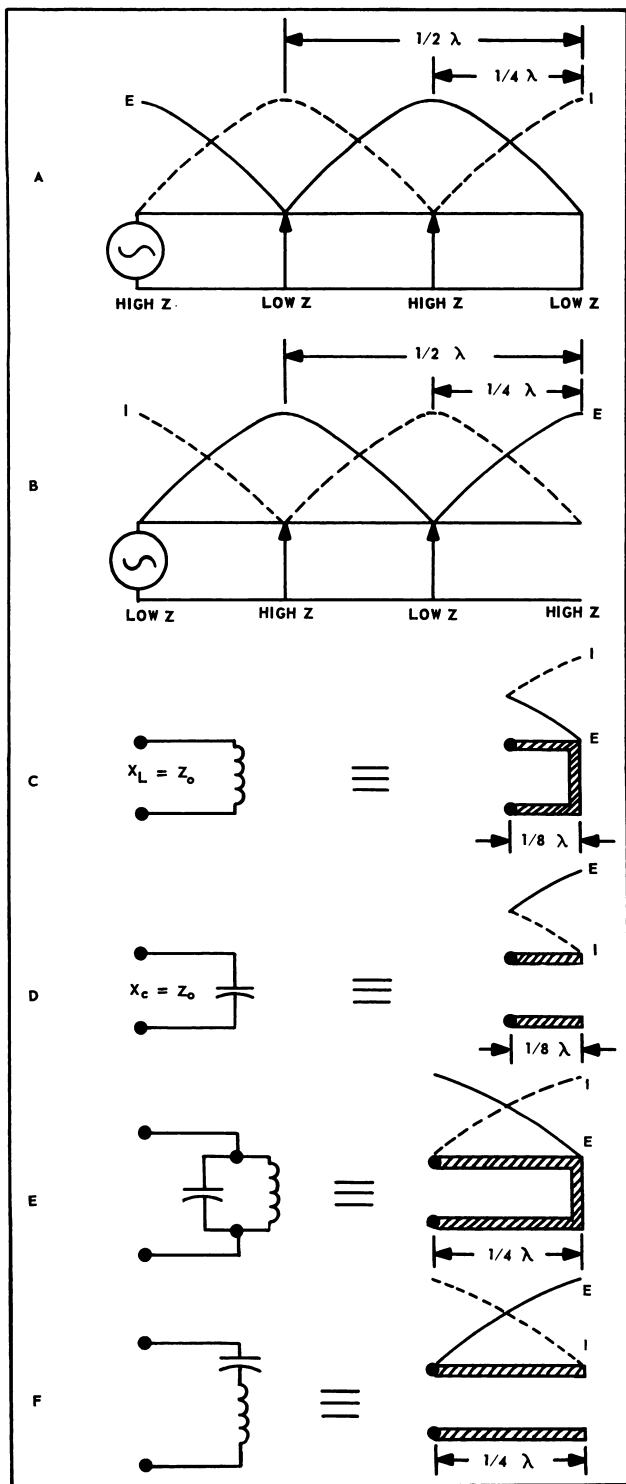


Figure 10-16. Sending End Impedance of Various Lengths and Terminations

quarter-wave intervals. In Figure 10-16B, where the line is open-circuited, the power factor is alternately capacitive and inductive at quarter-

wave intervals, exactly as in the case of the shorted line. However, the open line is capacitive over the quarter-wave section nearest the receiving end. These variations in the impedance of a line are shown at the top of Figure 10-20.

### Transmission Lines as Resonant Circuits

At the points of voltage maxima and minima on a short-circuited or open-circuited line, the line impedance is resistive. On a short-circuited line, the points an odd number of quarter-wavelengths from the receiving end have a high impedance (Figure 10-16A). If the frequency of the applied voltage to the line is varied, this impedance decreases as the effective length of the line changes. This variation is exactly the same as the change in the impedance of a parallel resonant circuit when the applied frequency is varied.

At the points an even number of quarter-wavelengths from the short circuit, the impedance is extremely low. When the frequency of the voltage applied to the line is varied, the impedance at these points increases in the same way as the impedance of a series resonant circuit varies as the frequency applied to it is changed. The same thing is true for an open-ended line (Figure 10-16B) except that the points of high and low impedance are reversed.

Since the impedance a generator sees at the quarter-wave point in a shorted line is that of a parallel resonant circuit, a shorted quarter-wavelength of line may be used as a parallel-resonant circuit (Figure 10-16E). An open quarter-wavelength of line may be used as a series-resonant circuit (Figure 10-16F). The Q of such a resonant line is much greater than can be obtained with lumped capacitance and inductance.

### Transmission Lines as Circuit Elements

A shorted section of line less than a quarter-wavelength appears inductive to a generator. If the line is exactly an eighth-wavelength, the inductive reactance is equal to the  $Z_0$  of the line. This is shown in Figure 10-16C. An open section of line less than a quarter-wavelength appears capacitive to a generator. If the line is an eighth-wavelength, the capacitive reactance is equal to the  $Z_0$  of the line. This section is shown in Figure 10-16D.

## Voltage, Current and Impedance Inversion

When the quarter-wave section is open at the end, the voltage at the end is high and the current is low, as you can see in Figure 10-16F. At the sending (generator) end the current is maximum and the voltage is low. Effectively the quarter-wave section has inverted the voltage and the current. The low voltage and high current at the generator indicates a low impedance. The quarter-wave section has inverted the impedance at the receiving end (open) to a short circuit at the sending end. Similarly, a shorted quarter-wave section will invert the short at the receiving end to an open at the sending end.

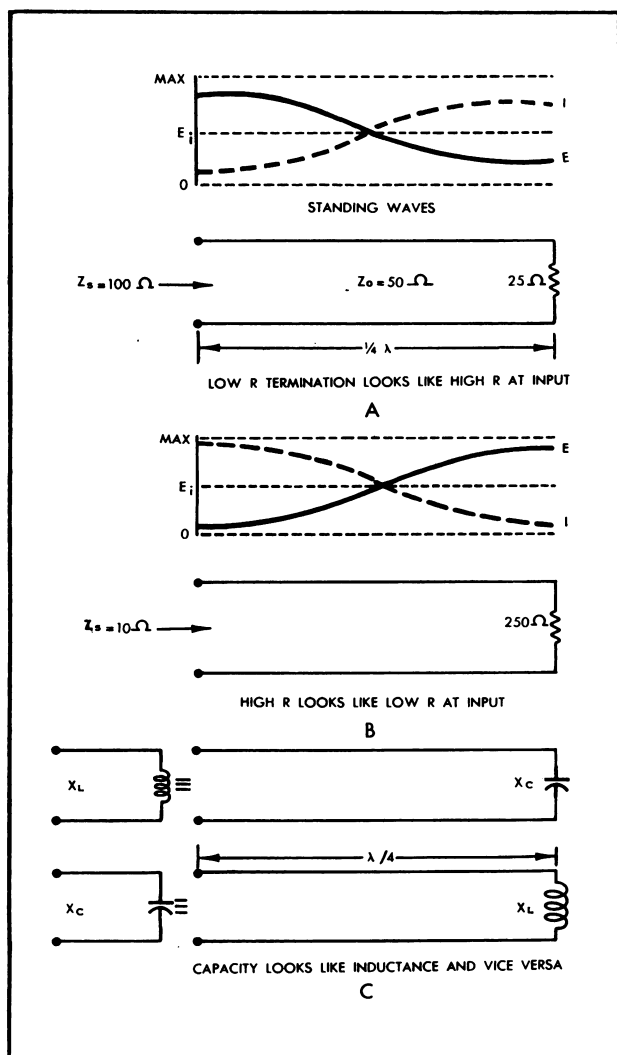


Figure 10-17. Impedance Inversion with Quarter-Wave Section

There are other terminations for the quarter-wave sections. When a quarter-wave section is terminated in a resistance greater than  $Z_0$ , the section will invert it to resemble a resistance less than  $Z_0$ . Mathematically, the inversion is expressed by the equation,

$$\frac{Z_R}{Z_s} = \frac{Z_0}{Z_s} \quad (19)$$

where  $Z_R$  is the impedance at receiving end,  $Z_s$  the impedance at sending end, and  $Z_0$  the characteristic impedance of the line.

For example, the impedance  $Z_s$  at the sending end of a 50-ohm line, terminated with a 25-ohm resistor, is equal to,

$$Z_s = \frac{Z_0^2}{Z_R} = \frac{50^2}{25} = \frac{2500}{25} = 100 \text{ ohms}$$

From equation (18) you can calculate the reflected voltage from this termination to be one-third of the incident voltage with a negative sign. The standing wave of voltage at the end is equal to two-thirds of the incident voltage. With this information you can determine the standing wave pattern to be as shown in Figure 10-17A. At the sending end the voltage is higher, but the current ratio at the sending end terminals is 100 (ohms).

If the terminating resistance is greater than  $Z_0$ , the input impedance will be lower than  $Z_0$ . For example, if the same line were terminated with 250 ohms, its input impedance would be equal to,

$$Z_s = \frac{Z_0^2}{Z_R} = \frac{50^2}{250} = \frac{2500}{250} = 10 \text{ ohms}$$

When the 250 ohms is inverted, it will look like 10 ohms, as illustrated in Figure 10-17B.

Effectively the quarter-wave section is an impedance-matching transformer at *one frequency*. This inverting property also holds for reactance. If the section is terminated in capacitive reactance, the input will be inductively reactive. If the section is terminated in inductance, the input will display all the characteristics of a capacitor, as illustrated in Figure 10-17C.

## Three-Eighth-Wave Section

When the r-f line is exactly three-eighths wave long, you can determine the input impedance as easily as you can with a quarter-wave section. For example, consider the shorted section shown in Figure 10-18A. This can be reduced, for

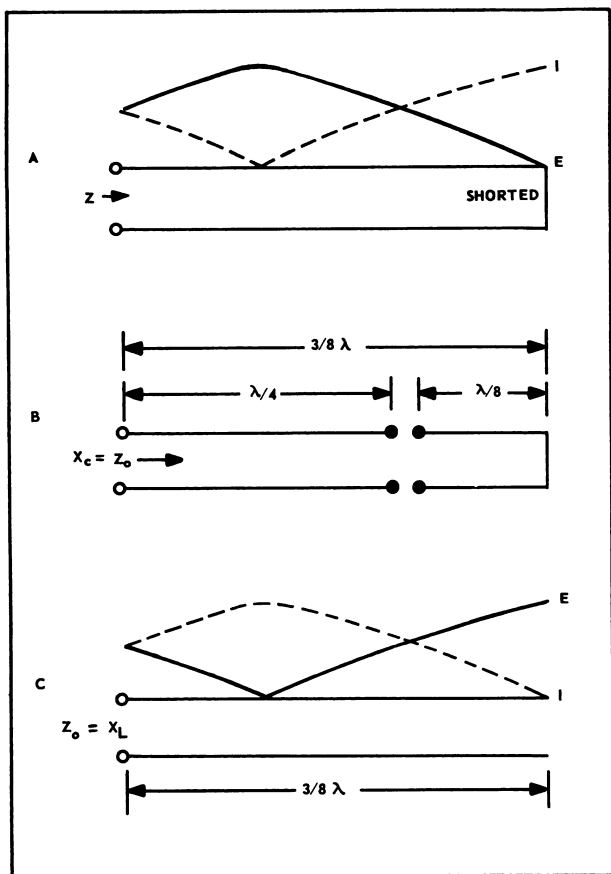


Figure 10-18. Three-Eighth-Wave Section

simplification, to two sections as shown at B. Here, an eighth-wave section is connected to a quarter-wave section. The eighth-wave shorted section displays inductive reactance to the quarter-wave section. Since a quarter-wave section inverts the inductance to look like capacitance, the three-eighth wave section is capacitive. The capacitive reactance is equal to  $Z_0$ . If the end is open as illustrated in Figure 10-18C, similar reasoning will show the line to be inductively reactive at the input terminals. Reference to the standing waves will verify this reasoning, for at the sending end you see the same situation as at the end of a line terminated in reactance—that is, the standing waves are between minimum and maximum.

### Half-Wave Section

When the section is a half-wave long, the situation resembles that of two quarter-wave sections connected together, Figure 10-19. The first quarter-wave section inverts the terminal

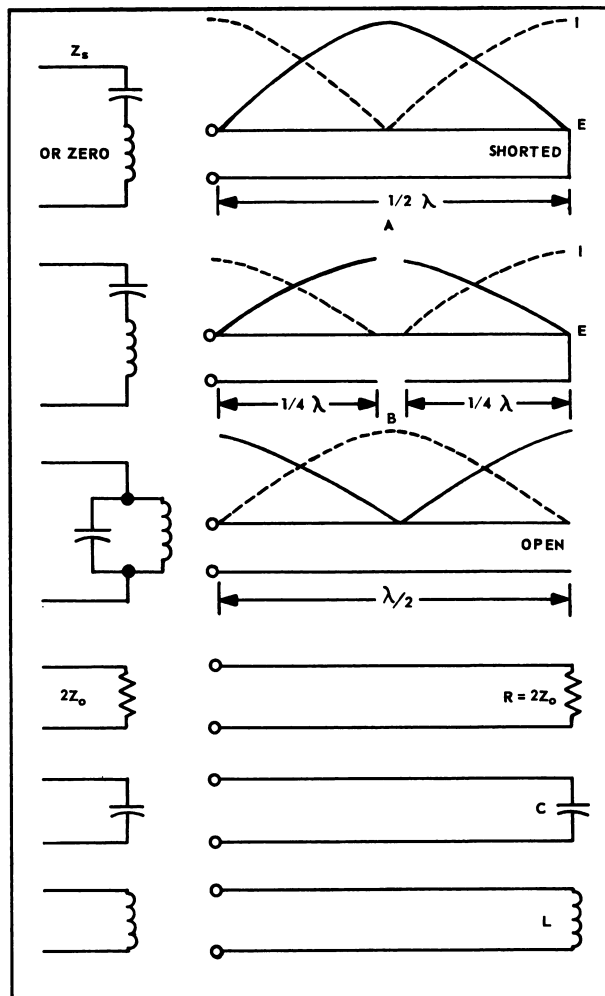


Figure 10-19. Half-Wave Section

impedance and the second inverts it again as shown in Figure 10-19B, making the input impedance the same as the terminating impedance as shown at A. The input impedance to a shorted half-wave section is zero impedance because the voltage is zero and the current maximum. In addition, the properties of the quarter-wave resonant circuit also are inverted.

For example, in Figure 10-18B, the shorted quarter-wave section is like a parallel resonant circuit. This is inverted, and looks like a series resonant circuit at the input to the half-wave section. In a like manner, the half-wave section, open at the end, has a high impedance at the input terminals and displays reactance to frequencies at which it is not a half-wave, so it is like a parallel resonant circuit. When the half-wave section is terminated in resistance (other



than  $Z_0$ ), in capacitance, or in inductance, it will always repeat this impedance at the sending end.

Lengths of line which are not exact multiples of those discussed will display reactance if terminated in a short circuit or an open circuit, but by varying the length and termination, they can be made to display resistance. If the eighth-wave section is shortened, it becomes less than an eighth-wavelength, and it will still be inductively reactive, but by an amount less than  $Z_0$ . If it is longer than an eighth-wavelength, it will display an inductive reactance greater than  $Z_0$ .

### Impedance Chart for Open and Shorted Lines

Figure 10-20 shows the relation of voltage, current, and impedance for various lengths of both open-end and closed-end transmission lines. Directly above the line is shown the impedance which the generator sees for various lengths of line. The curves above the letters of various height indicate the relative values of the impedance presented to the generator as it moves

from right to left. The circuit symbols indicate what the equivalent electrical circuits are for the transmission lines at that particular length. The standing waves of voltage  $E$  and current  $I$ , whose ratio is the impedance  $Z$ , are shown without polarity above each line.

### TYPES OF R-F LINES

The principal types of r-f lines are the twisted two-wire line, the spaced two-wire line, the shielded two-wire line, the flexible coaxial cable, and the rigid coaxial cable.

#### Twisted Two-Wire Line

The twisted two-wire line in Figure 10-21 consists of a pair of insulated and twisted conductors. The twisting holds the lines together mechanically and aids in balancing each conductor to the influence of nearby magnetic and electrostatic fields. Twisted wires usually have a characteristic impedance of 70 to 100 ohms. Although the

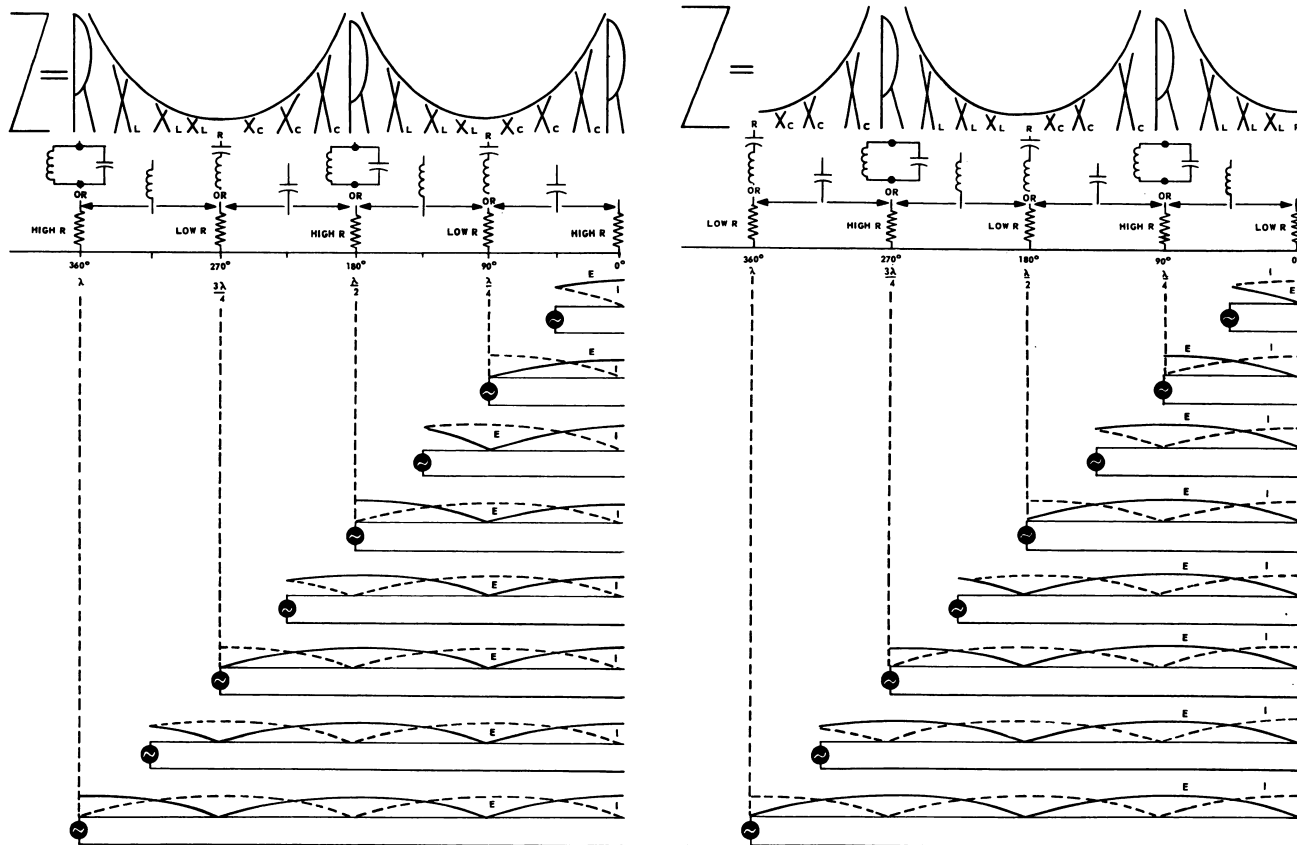


Figure 10-20. Impedance Charts



Figure 10-21. Twisted Two-Wire Line

twisted line is flexible, occupies little space, and is simple to manufacture, it is not usable at radar frequencies because of capacitance between the wires, lack of shielding, and changes in characteristics due to moisture on the line.

### Spaced Two-Wire Line

The spaced two-wire line (Figure 10-22) consists of two parallel conductors that are maintained a fixed distance apart by insulating spacers or spreaders. This line has two types—the spreader bar, and the newer twin-lead line. The spreader bar type uses ceramic or polystyrene bars for spacing wires. The spacing may be up to six inches or more. The twin-lead line is the result of improved insulating materials and manufacturing techniques. It consists of two wires which are molded into a low-loss plastic called polyethylene. Impedances ranging from 75 to 300 ohms are available in the twin-lead line.

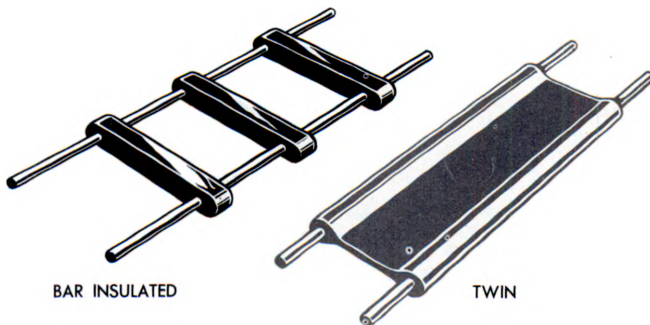


Figure 10-22. Spaced Two-Wire Line

Lines in general, however, may be constructed with any impedance up to 700 ohms by varying the wire diameter and spacing. The twin-lead line has a number of advantages over the spreader bar two-wire line. Wide spacing between the leads in the twin-lead type gives lower shunt capacitance between the leads. Insulation losses are lower in it, since the wires are separated by

something other than air, unlike the wires in the spreader bar line.

The twin-lead two-wire line is not perfect. It has two principal disadvantages. In the first place the line acts as an antenna. The magnetic fields extend some distance out from the conducting wires. When these fields collapse at a change in frequency cycle, some of the lines of force do not retract when the change is rapid, and consequently a few lines of force are projected outward in the form of radiation, just like radiated energy on an antenna. This energy is lost. In the second place, nearby objects set up capacitance with the wires and disturb current flow through the line. This obstructs the orderly transfer of energy by the line.

### Characteristic Impedance of the Two-Wire Line

The relationship between the characteristic impedance and the dimensions of a two-wire line is expressed mathematically by the equation,

$$Z_o = 276 \log_{10} \frac{2D}{d} \quad (20)$$

Where  $D$  is the distance between wires (center to center), and  $d$  the diameter of one wire.

#### Example

##### Problem

Find the  $Z_o$  of a two-wire line constructed of No. 12 wire (.08 inch in diameter) spaced six inches apart.

##### Solution

$$\begin{aligned} Z_o &= 276 \log_{10} \frac{2D}{d} \\ &= 276 \log_{10} \frac{2 \times 6}{.08} \\ &= 276 \log_{10} 150 \\ &= 276 \times 2.1761 \end{aligned}$$

The characteristic impedance ( $Z_o$ ) then, is approximately 600 ohms.

The graphs shown in Figure 10-23 are for plotting characteristic impedances. One is for a two-wire line. The other is for coaxial lines, which are explained later. In Figure 10-23A, dotted lines show how the characteristic impedance can be obtained from the impedance curve.

### Shielded Two-Wire Line

A shielded two-wire line is similar to the two-wire spaced line except that it is shielded by a copper braid. Notice the practical form of this line in Figure 10-24. The copper braid gives it



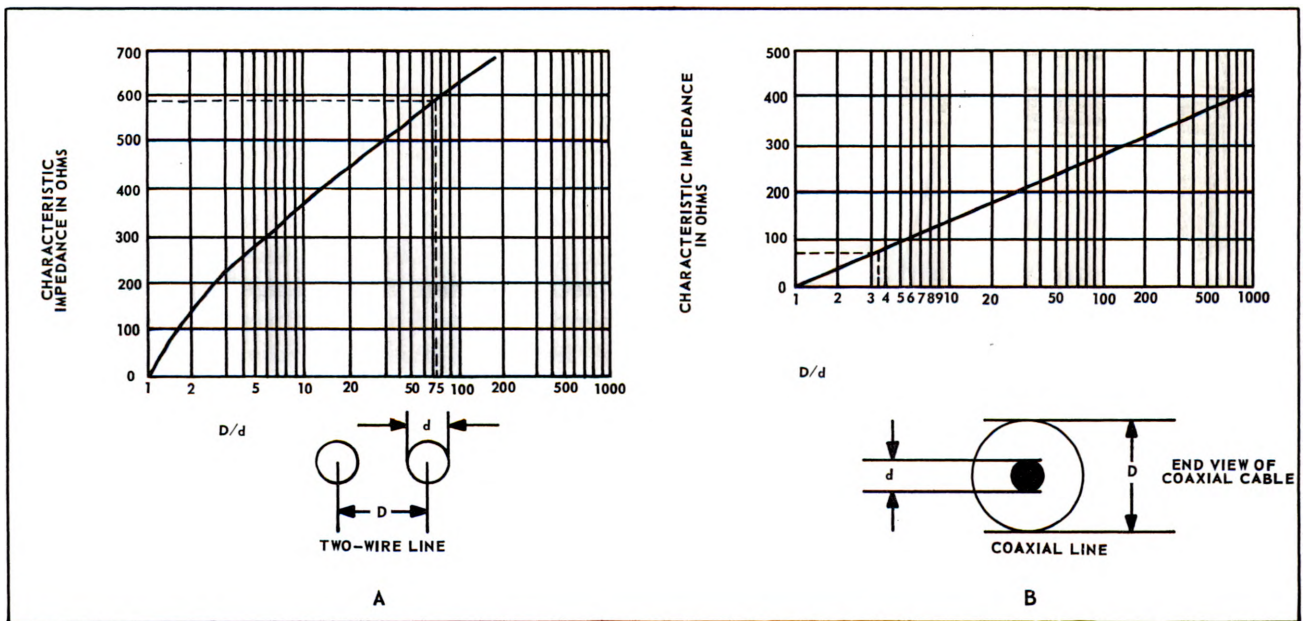


Figure 10-23. Characteristic Impedance Graphs for Two-Wire and Coaxial Lines

flexibility. Flexibility could be obtained by using unbraided tubing. However, the use of braid permits uniform spacing of each conductor during manufacture. This results in each wire being perfectly balanced capacitively to the surrounding conductor. As long as the balance is maintained, certain detrimental effects, such as high capacity to ground when the shield is grounded, are very slight. This line does not radiate energy because of the shield and thereby is not affected by nearby magnetic fields.

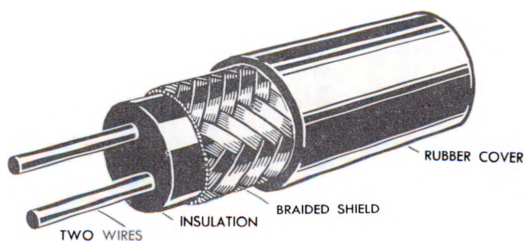


Figure 10-24. Shielded Two-Wire Line

### Coaxial Line

The coaxial line is the most universally used type of line for r-f transmission. It consists of two conductors. One of the conductors is hollow; the other conductor is located inside the hollow conductor. The two conductors are made con-

centric in cross section to maintain uniform characteristics. Early coaxial lines were made of copper tubing in which there was a wire held in the center by ceramic or polystyrene washers or beads. Attempts to build flexible cables led to the use of outer conductors of copper braid spaced by a continuous row of beads or rubber insulators. But rubber insulation caused excessive losses at high frequencies. In addition the bead arrangement permitted air to enter the line, causing moisture to collect. This resulted in high leakage currents and arc-over when high voltages were applied.

Research solved this problem by using polyethylene plastic, a hard, solid material, safely flexible over a temperature range of  $-40^{\circ}$  to  $+80^{\circ}$  C. Polyethylene is unaffected by such fluids as acids, alkalis, aviation gasoline, oil, hydraulic brake fluid, and sea water. Furthermore, there is no known solvent for polyethylene at ordinary temperatures. Its extremely low losses at high frequencies is important for radar. At 200 mc the losses may be only one-hundredth of those of rubber under certain conditions. When carefully manufactured, a coaxial cable insulated with polyethylene will be uniform within .005 inch and will have no air between the conductors. Thus, due to its uniform characteristics, it will operate indefinitely with low leakage and with little danger of arc-over.



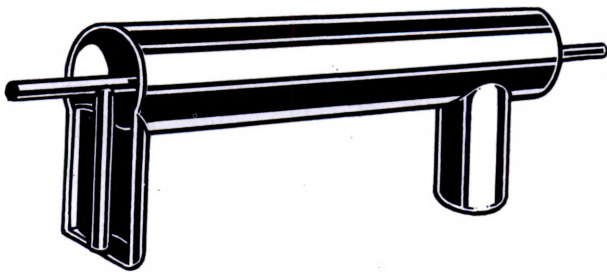


Figure 10-25. Rigid Coaxial Cable with Metallic Insulators

Where installations require still lower losses, such as 3000-mc airborne radar installations, an air-insulated rigid coaxial cable meets these requirements. In this cable (Figure 10-25) the center conductor is supported with metallic insulators which are quarter-wave sections of coaxial line. Since there is air in the interior, the interior of the line is usually pressurized to keep moisture out.

Coaxial lines have a number of advantages. The shielding is perfect for both magnetic and electrostatic fields. If you refer to the two end views of the coaxial cable in Figure 10-26, you can see that the electrostatic field is terminated at the outer conductor so that none of the field is outside of the line. The magnetic field from the inner conductor extends beyond the outer conductor, but an equal amount of current flows in the outer conductor, setting up the field shown, which is outside of the outer conductor. These two fields cancel. Inside the cable these two fields add. Hence, there is neither an electrostatic nor a magnetic field outside of the cable. For the same reasons that a coaxial line does **not radiate**, it also does not pick up any energy.

Therefore, it can be installed anywhere without being influenced by other strong fields. Also, its capacitance is lower than that of a shielded pair line.

Some of the disadvantages of coaxial lines are: they cost more than two-wire lines, they are more difficult to install than flexible cables, and it is more difficult to measure the fields in a coaxial cable than on an open-wire line.

#### Characteristic Impedance of Coaxial Lines From Dimensions

When you know the dimensions of a coaxial line, you can determine its characteristic impedance by the following equation,

$$Z_0 = 138 \log_{10} \frac{D}{d} \quad (21)$$

where  $D$  is the inside diameter of the outer conductor and  $d$  is the diameter of the inner conductor.

#### Example Problem

If the inner conductor is a quarter-inch rod, and the outer conductor is a 0.875-inch tubing (inside diameter), find  $Z_0$ .

#### Solution

$$\begin{aligned} Z_0 &= 138 \log_{10} \frac{D}{d} \\ &= 138 \log_{10} \frac{0.875}{0.25} \\ &= 138 \log_{10} 3.5 \\ &= 138 \times 0.5428 \end{aligned}$$

The characteristic impedance ( $Z_0$ ) equals 75 ohms.

You can conveniently plot this equation in the impedance vs. dimensions graph shown in Figure 10-23B. The dotted lines show how  $Z_0$  for this example is obtained graphically. A popular value for characteristic impedance is 75 ohms since the

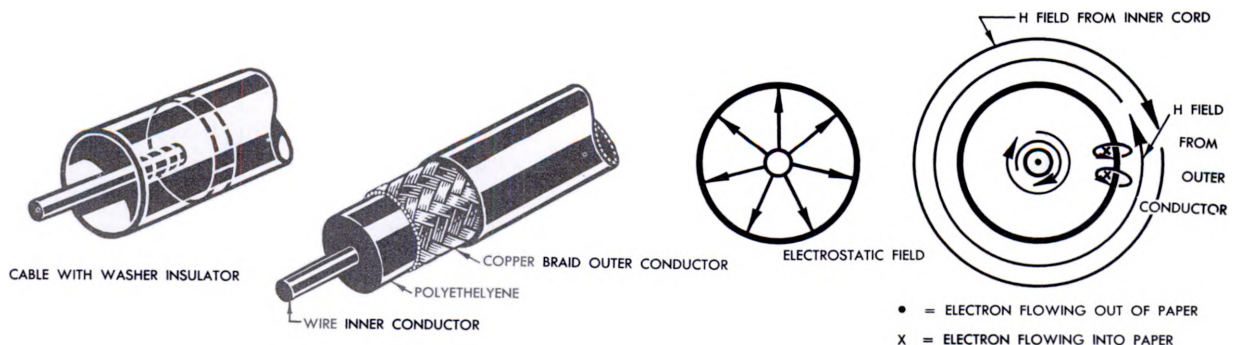


Figure 10-26. Coaxial Cables

Q of a coaxial line is highest when its dimensions provide for 75 ohms resistance.

## LOSSES IN R-F LINES

Although the theory of r-f lines just given dealt with lines assumed to be lossless, actually there are some losses in all lines. Line losses may be of three types — copper losses, dielectric losses, and radiation or induction losses.

### Copper Losses

One type of copper loss is  $I^2R$  loss. In r-f lines the resistance of the conductors is never equal to zero. Any time current flows through one of these conductors, there is some energy dissipated in the form of heat. This loss of energy is an  $I^2R$ , or power loss, since the formula for power is  $P = I^2R$ . With copper braid, in which the resistance is higher than the resistance in a solid tube, this power loss is accordingly higher.

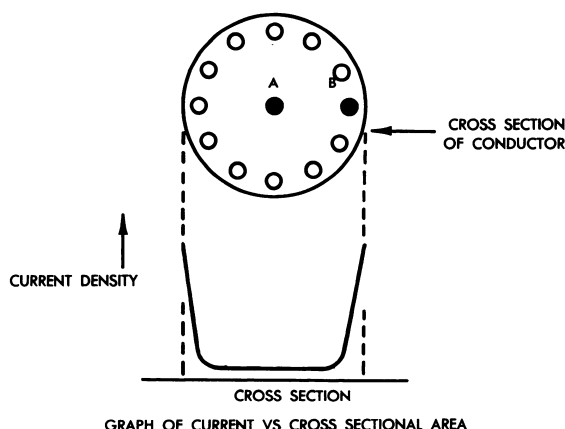


Figure 10-27. Skin Effect Losses in R-F Lines

Another copper loss results from skin effect, which can be explained as follows: When a direct current flows through a conductor, electrons move uniformly through the cross section of the conductor. On the other hand, when an alternating current flows, this is not the case. Expanding and contracting magnetic fields about each electron encircle other electrons and retard their movement. This is the phenomenon of self-induction. From Figure 10-27, you can see that the fields about the electrons moving in area A

expand and affect the movement of all the electrons near the surface of the conductor at B. At the same time all the electrons moving at the surface have equal fields. These fields expand and retard the electrons at A.

While a single electron exerts only a small effect on the outer electrons, the outer electrons in combination exert a tremendous effect on the center electron — so much so, that at frequencies of 100 mc and above, electron movement (current flow) is so small at the center that the center of the conductor can be removed without noticeable effect on current flow. The cross section (because current flows only in the outer portion of the conductor) is less and the actual resistance is greater, since the resistance of a conductor varies inversely with the cross section.

In conclusion, skin effect is the tendency of alternating currents to flow near the surface of the conductor, thus being restricted to a small part of the total cross-sectional area. This effect increases the resistance and becomes more marked as the frequency rises.

Because of the flow of current near the surface of a conductor, the conductivity of an r-f line can be increased by plating it with silver. Most of the current will flow in the silver layer. The tubing serves chiefly for mechanical support.

### Dielectric Losses

Dielectric losses are due to the heating of the dielectric material (insulation) between conductors. This heating of the dielectric material takes power from the source. The heat itself is dissipated into the surrounding medium. Figure 10-28 shows the mechanics of dielectric losses. When there is no potential difference between two conductors, the atoms in the dielectric material between them are normal; that is, the orbits of the electrons are circular, as shown in Figure 10-28A. But, when a potential difference exists, as shown in Figure 10-28B, the electron orbits change.

The excessive negative charge on one conductor repels electrons in the dielectric toward the positive conductor and thereby distorts their orbits. This change in the path of electrons requires work. The power is furnished from the power source for the r-f line, thus introducing a power loss. The structure of the atoms in some dielectric materials, rubber for example, is harder to distort than the structure of other materials. On

the other hand, the atoms of some materials distort easily. Polyethylene is this type of material. It is used as a dielectric because little power is consumed when its electron orbits are distorted.

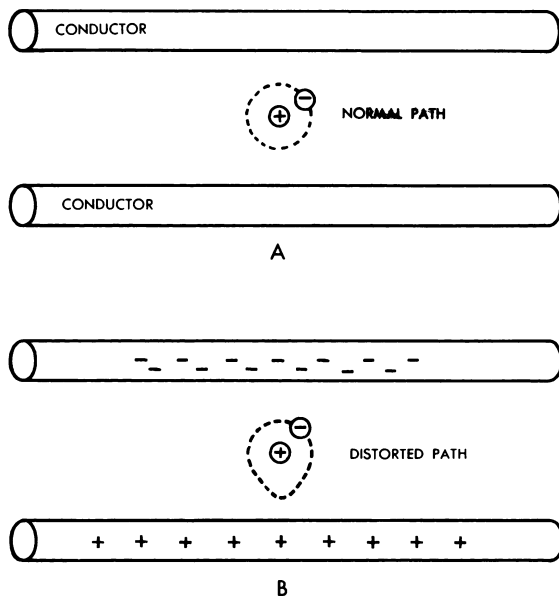


Figure 10-28. Origin of Dielectric Losses

### Radiation and Induction Losses

Radiation and induction losses are similar in that they are both due to the fields surrounding conductors. When the field about conductors cuts a nearby metallic object, there is a current induced in the object. This induced current results in power being dissipated by the object. The power thus lost is supplied by transformer action from the source for the r-f line. Radiation losses occur because some lines of force about a conductor do not return to it when the frequency cycle changes. These lines of force are projected into space as radiation, and since they do not return, the energy they use must be supplied by the power source, thus introducing additional power loss.

### Importance of Losses

Although in cross-country telephone lines resistance is a major problem from the standpoint of losses, resistance is of little concern in r-f systems, since r-f lines are short and the DC resistance is quite small. However, the other types of losses previously discussed are of more

concern. Most copper losses in r-f lines are due to skin effect. These, however, can be minimized by using large diameter conductors with a low resistance surface material.

Dielectric losses, on the other hand, are of considerable concern. To obtain minimum dielectric loss, it is necessary to use *dry air* as the dielectric. When the line is to be flexible, it is necessary to use a low loss dielectric, such as polyethylene. Furthermore, since dielectric losses increase with the length of the line, systems requiring extremely low loss use a line as short as possible.

In two-wire lines, radiation losses can be decreased either by reducing the spacing between wires or by shielding. Both methods increase the capacitance between the wires, which is usually undesirable. For the least losses from this cause, coaxial cables are recommended because of their natural shielding. This type of loss increases as the square of the frequency. Thus, two-wire lines are not usable at 3000 mc and up. Shielded and coaxial cables are used for high-frequency radar systems.

### MEASUREMENTS IN R-F LINES

Electrical measurements must be made on the r-f lines to be certain that they are transferring energy with the least possible losses, to find the electrical length, and to obtain other pertinent data.

There are several simple methods for determining the magnitude of voltage or current at any point on an r-f line. In using these methods remember that the magnetic field around a line varies directly with the current, and that the electrostatic field about the line varies directly with the voltage.

One method for observing current at any point on a line is to break the line and insert an ammeter. Figure 10-29A shows an AC ammeter connected in series with the line. Here the meter reads the standing wave of current at the point indicated. Sometimes it is not practical to cut a line. In that case a magnetic loop that is connected to a meter is moved along the line. With this method the meter reads maximum at the point where the greatest current is induced in the loop — that is, at the current loop — and it reads minimum at the nodes, or point of current minimum.

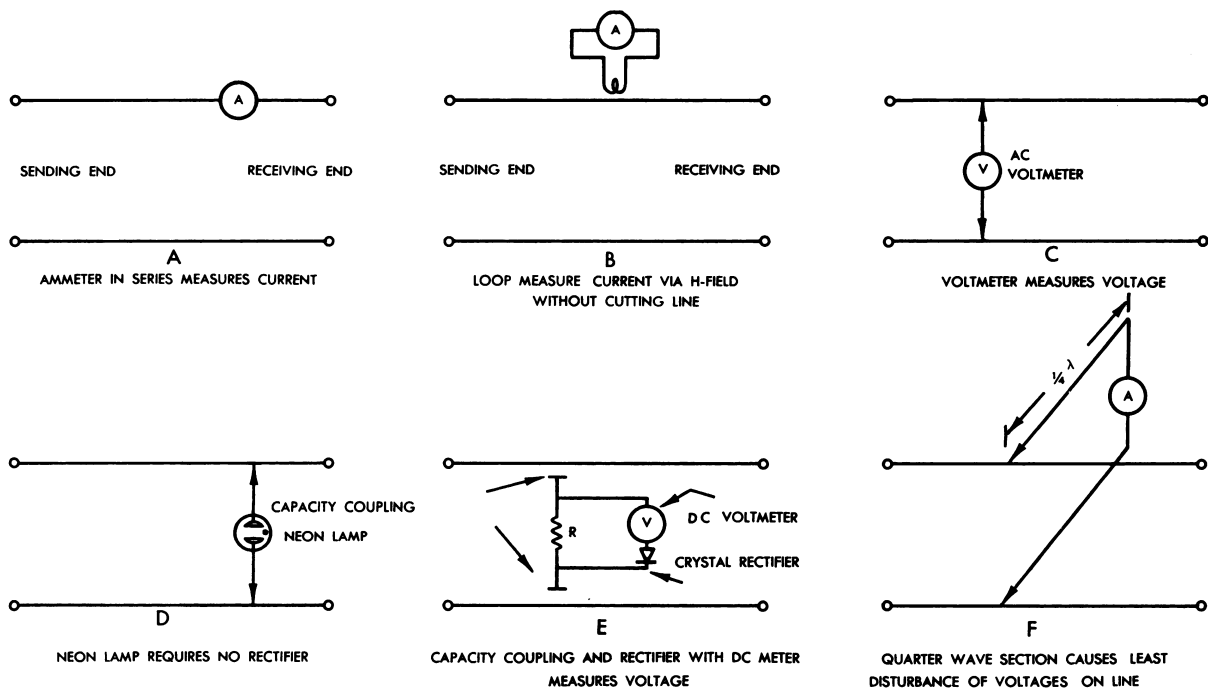


Figure 10-29. Making Measurements on R-F Lines

To read voltage at any point, connect an AC voltmeter across the line at the point you are checking as shown in Figure 10-29C.

Another simple method is to connect a neon lamp across the line. When the lamp is connected across the line, the voltage in the line causes the lamp to glow. The greater the voltage, the brighter the lamp will glow. A strong enough electrostatic field will make a neon lamp glow, even when the lamp is not directly connected to the line. In high-powered installations you need only to place the neon lamp near the line. A more accurate device is the sensitive DC meter with a rectifier. A crystal or diode may be used as the rectifier, as shown in Figure 10-29E. The resistor shown is capacitively coupled to both wires; the voltage drop across the resistor is measured by the meter.

All the preceding methods of measuring have the common disadvantage that ordinary measuring devices take a certain amount of energy from the line being measured. Such use of energy represents a lowered impedance at the point of measurement. Any change in impedance at any point along a line causes reflections. Thus, ordinary measuring instruments load (take energy from) a line and upset its normal characteristics.

A high impedance measuring device suitable for making measurements in a two-wire line is the quarter-wave section shown in Figure 10-29F. This device, which consists of an AC ammeter, forms a short circuit across the section. As the quarter-wave section is shorted at the ammeter end, it presents an extremely high impedance to the line and very little current is needed to energize it.

When coaxial lines are used, as is the case with radar equipment, the magnetic and electrostatic fields are not accessible to loops and contacts. In measuring the field in this type of line it is necessary to use a slotted line section as in Figure 10-30. A probe is inserted in the slot in the line not far enough to touch the center conductor. The probe is a slender rod that acts as an antenna and is excited by the electrostatic field which is parallel to it. It is not affected by the magnetic field. The slot does not reduce the effectiveness of the coaxial line because the currents flow parallel to the slot. With this arrangement the coupling is very slight and therefore very little energy is extracted by the probe.

The energy at the probe is AC at a radio frequency and is detected with some form of detector. A vacuum tube detector can be used, but usually a crystal detector or a bolometer is used.

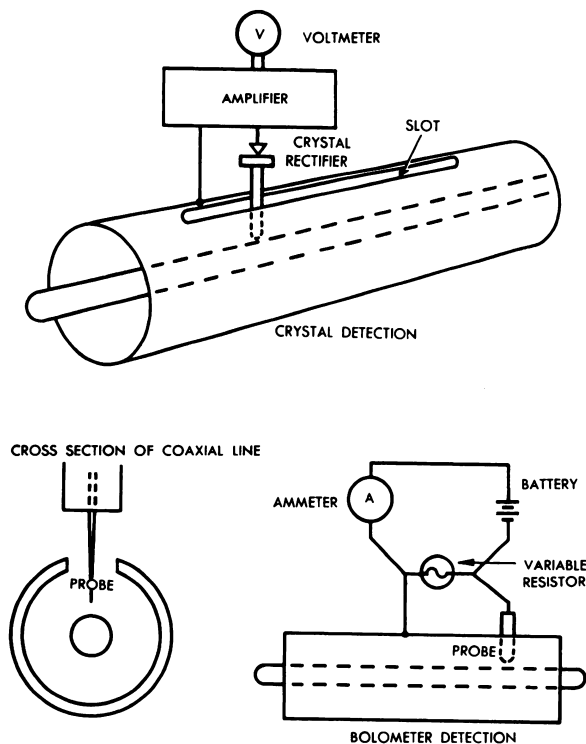


Figure 10-30. Crystal and Bolometer Detection

In the crystal detector, a crystal rectifier changes the r-f to DC, whose magnitude is determined by the amplitude of the AC voltage at the point where the probe is inserted. The DC can be connected to a DC meter. Since the DC is usually quite small, it is first amplified by a vacuum tube amplifier before it is fed to the meter.

A bolometer is a piece of very fine wire in which the resistance varies with temperature. As it is used in Figure 10-30, the r-f energy heats the wire. The heat changes the resistance of the wire to the direct current flowing from the battery. The meter then changes its reading. This change also can be amplified by a vacuum tube amplifier. The resistance element in the bolometer is usually a standard 1/100 ampere fuse.

### THE LECHER LINE

*Lecher line* is the term applied to a two-wire r-f line which is one-quarter to two or three wavelengths long. Such a device is used for either investigating standing wave phenomena, meas-

uring half-wavelengths, or utilizing the principles of an r-f line. In the Lecher line in Figure 10-31A, two wires run parallel for about one wavelength. With this line you can determine several facts by experimentation. For example, when the frequency of a transmitter is unknown, you can use a Lecher line to measure the wavelength of the transmitted frequency. Further, since the velocity at which energy travels on a line of this type is about 97.5% of the velocity in free space (300 million meters per second), you can compute the frequency by the formula,

$$f = \frac{300 \times 10^6 \times 0.975}{2\omega} \quad (22)$$

where  $\omega$  is the distance between two minimum or maximum standing wave points, and  $f$  the frequency in cycles per second.

Generally, a sliding shorting bar is a part of a Lecher line. As the sliding bar moves toward the generator, the standing waves move with the shorting bar. Results are read by a pickup coil with an indicating meter which is placed near the excitation coil of the Lecher line.

The accompanying illustration explains the use of the shorting bar and the pickup coil in determining current minimums and maximums. The standing current wave shown in Figure 10-31A shows that the standing wave of current is minimum at the location of the excitation coil. Thus, very little current flows through it, and the small magnetic field resulting induces little voltage in the meter circuit. The meter will indicate a current minimum. In Figure 10-31B, which shows the shorting bar moved, the current is maximum at the excitation coil, and the meter pickup coil is excited by the strong magnetic field about the excitation coil. The high current on the meter shows the presence of a current maximum.

A Lecher line one-quarter wavelength long displays the characteristics of a parallel resonant circuit. For this reason, a quarter-wave Lecher line often is used as a tuned circuit in an ultra high-frequency oscillator. At 200 mc a quarter-wave is only about 14 inches, which is small enough for practical use as is shown in Figure 10-31C. (Note the typical Lecher line oscillator at D.) A Lecher line oscillator is impractical at lower frequencies. For example, at a frequency of one mc, a quarter-wave line is 250 feet long. Obviously, this is too long for the ordinary transmitter operating at these frequencies.



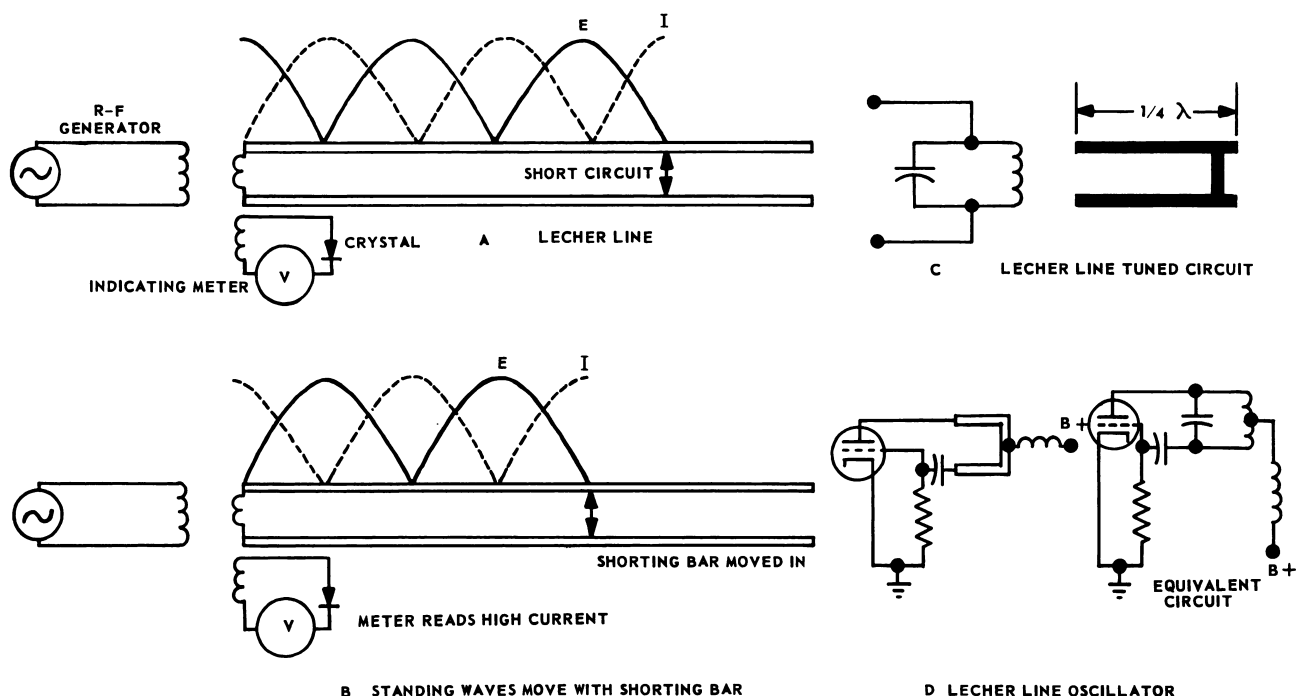


Figure 10-31. Use of Lecher Lines

### USE OF R-F LINES

In radar, standing-wave ratios must be as near unity as possible for satisfactory operation. However, even with a near unity ratio, standing waves cause several deficiencies. First, they reduce power handling capacity. The maximum power that a line can handle is limited by the maximum voltage that can be impressed across the line. The maximum voltage, in turn, is limited by the breakdown voltage of the insulation. Standing waves increase the voltage at some points without additional power. Second, the efficiency of the line is lower with standing waves because of the high circulating currents at the current loops. The  $I^2R$  losses introduced by this current cannot be tolerated. Third, standing waves change the input impedance of the line. The frequency and power output of the usual radar transmitter varies with load impedance, so the line can change both the transmitter frequency and power output. In practice it is usually possible to get the standing-wave ratio down to less than 1.5:1, a ratio which is considered satisfactory.

When energy is transferred from the antenna to the receiver, it is also desirable to get a low standing-wave ratio, since any losses that occur

subtract from the energy received by the antenna system. A low standing-wave ratio also prevents excessive attenuation of the received echoes, which are usually too weak to stand much loss.

### Metallic Insulators

In any of the previous uses of r-f lines described, standing waves should not exist because of the unusual phenomena which result. However, there are many uses of r-f lines other than the transfer of energy in which standing waves are desirable. The metallic insulator is an example (Figure 10-32). Since radar sets operate at a single frequency, it is possible to support the center conductor of the coaxial line by a quarter-wave section of r-f line.

The inverting properties of this quarter-wave section make it possible to short the receiving end of the section. This makes the input ends look like an infinitely high resistance. However, because of slight copper losses, the infinite resistance is not quite attained, but the input resistance nevertheless is very high. The metallic insulator has no dielectric losses and is not damaged by arc-overs. Ordinary insulating materials are carbonized by an arc. The carbonized path

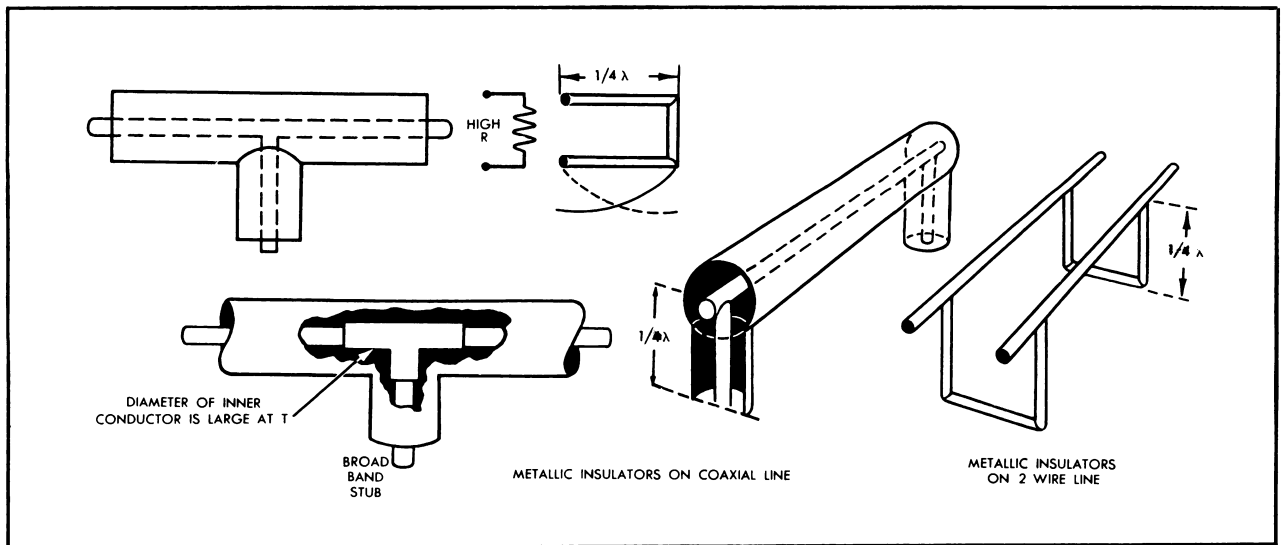


Figure 10-32. R-F Line Used as Metallic Insulator

acts as a low resistance conductor at any voltage. Air insulation in quarter-wave sections does not carbonize, and the line is not shorted after the arc.

Metallic insulators are practical only at the higher frequencies, such as 3000 mc, where the quarter-wave "stub" is about two inches long. At lower frequencies the physical length of the insulator would be too great for practical use. A disadvantage of this insulator is that it is a quarter-wave at only one frequency. This limits its use to constant-frequency equipment.

This is the function of the broad band stub. In pulse modulation, many sideband frequencies are present in a radar wave. At frequencies slightly different from the radar frequency a stub is normally reactive. This causes reflections in the main line. These reflections can be minimized by using a broad (frequency) band stub, like the one shown in Figure 10-32. Note that the diameter of the center conductor is enlarged at the junction.

### Impedance-Matching Device

The impedance of a shorted r-f line varies widely over its length. Figure 10-33A shows the variation of impedance along the length of a quarter-wave shorted r-f line. The impedance varies from zero at the short to an infinite value at a distance of a quarter-wave from the short. The curve shown is a hyperbola. When this section of r-f line is excited, it is possible to connect

devices of any impedance to it with perfect match somewhere along its length.

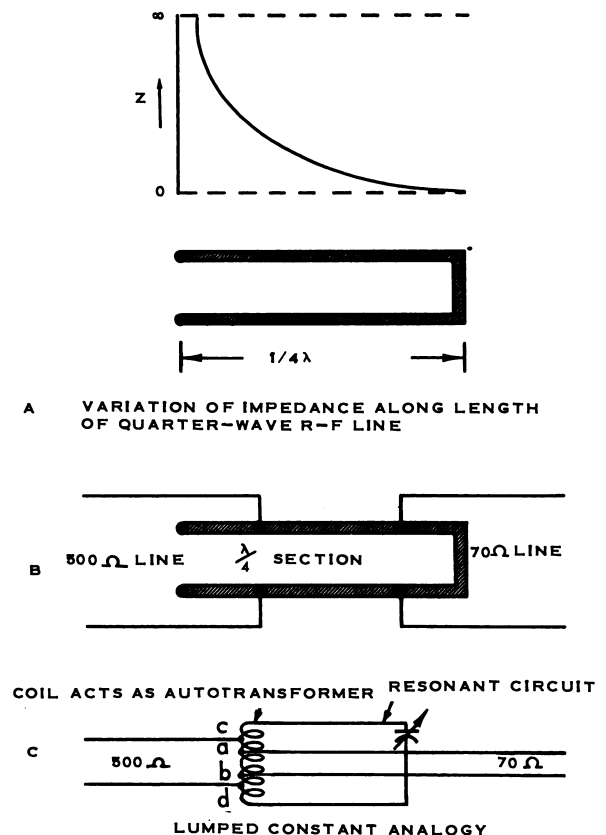


Figure 10-33. R-F Line Used as Impedance-Matching Device

For example, you can match both a 500-ohm line and a 70-ohm line to this section so that there will be no standing waves on either line. Figure 10-33B shows both the 500-ohm and the 70-ohm line connected to a quarter-wave section. Energy from the 500-ohm line sets up standing waves on the section. When the section is excited, the voltage-to-current ratio is correct at the point indicated. This means that the 500-ohm line is terminated in its  $Z_0$ . At the shorted end of the section, the voltage-to-current ratio is less. Therefore, at some point near this end of the section the value of  $E/I$  must equal 70 ohms. At this point you can properly match the 70-ohm line.

The action involved in matching a 500-ohm line to a 70-ohm line by a quarter-wave section is comparable to the action involved in a tuned autotransformer. Notice the diagram in Figure 10-33C. The action of the coil in this diagram is like that of an autotransformer. When a 70-ohm line is connected to points a and b, the reflected impedance between c and d is equal to 500 ohms. As in transformer action, the direction of energy transfer may be reversed without changing connections. The 70-ohm line can feed the quarter-wave section—a coil in this case—and the 500-ohm line will present the proper impedance to it. Any other value of impedance can be found for either line by moving the connections along the quarter-wave section.

Figure 10-34A shows still another way to get impedance-matching by using a quarter-wave section. Since a quarter-wave section will invert any value of impedance, by selecting a section of the correct impedance, you can invert the 70-ohm impedance of the line used in the preceding example to appear as 500 ohms at the other end. The equation for this relationship is

$$\frac{Z_s}{Z_0} = \frac{Z_0}{Z_r}$$

Solving this equation for  $Z_0$ ,

$$Z_0 = \sqrt{Z_s Z_r} \quad (23)$$

the impedance of the quarter-wave in Figure 10-34 is

$$Z_0 = \sqrt{Z_s Z_r} = \sqrt{500 \times 70} = \sqrt{35,000} = 187 \text{ ohms}$$

When you choose conductors of the proper dimensions and space them correctly, you can construct a 187-ohm line. In this arrangement there are standing waves on the quarter-wave section, but none on either line connected to it.

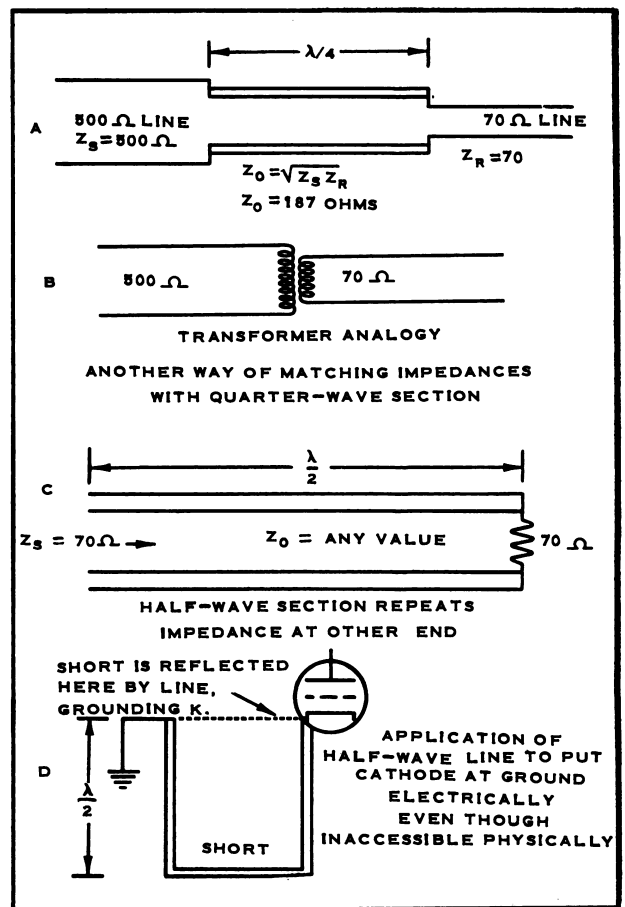


Figure 10-34. R-F Lines as Transformer

In the half-wave section in Figure 10-34C, the output impedance is repeated at the input. For example, a 70-ohm load is inverted over a quarter-wavelength to some other value, depending on the characteristic impedance of the line. It is further reinverted to 70 ohms over the remaining quarter-wave section. The entire action is like that of a one-to-one transformer. This type of inversion is frequently used in radar.

Sometimes a specific impedance is required between two tube elements, and connecting the impedance outside the tube does not provide the desired results. The connecting leads may be electrically long at the operating frequency. A half-wave line using the tube leads as part of the line will effectively place the required impedance right at the tube elements themselves. Figure 10-34D shows a half-wave line being used to put the cathode electrically at ground potential, even though it is not possible to physically place the ground within the tube.

Impedance-matching lines are not used solely with transmission lines. There are several types of devices that can be connected to an impedance-matching line. For example, the input to, or load on, an impedance-matching line could be an r-f device such as a transmitter, an antenna, a receiver input network, or a vacuum tube plate or grid circuit.

### Tuned Circuits

An r-f line having a certain length displays the properties of a resonant circuit. There are many cases in high-frequency equipment where this property is used. For example, quarter-wave and half-wave sections are commonly used as tuned circuits in radar equipment. The Hartley oscillator shown in Figure 10-35 uses a quarter-wave Lecher line for the tuned circuit. Compare this circuit with its equivalent circuit and note how the radar circuits differ from the conventional representation used in the equivalent circuit.

Another use of a section of r-f line is as a wavemeter. Figure 10-35 shows the r-f wavemeter. The meter consists principally of a coaxial line with a movable short-circuiting plate. When this plate is at a distance any multiple of a half-wavelength from the probe in the meter, the voltage at the probe drops to zero. The amount of voltage at the probe can be read on a rectifier type voltmeter. The distance represented by any two consecutive zero voltage readings on this meter represents one wavelength.

The knob at the right of the meter provides a means for moving the position of the short-circuiting plate. The distance traveled by this plate can be read from the metric scale, which is calibrated in frequencies.

### Phase Inverting or Shifting Networks

Sometimes it is desirable in electronic operations to shift the phase of current or voltage a certain number of degrees, or in some cases to invert the phase completely. Previously it was explained that the phase of voltage changes  $360^\circ$  in one wavelength of an r-f line. Thus it follows that the phase relationship of a sine wave at a distance of a half-wavelength from the sending end will be  $180^\circ$  out of phase with the sine wave at the sending end. This is the same action that occurs in a transformer. An RL circuit as shown in Figure 10-36, however, will shift the phase less than  $90^\circ$ . Lines made up of small parts of a wavelength shift the phase only in small amounts.

An example of a use of phase shifting is antenna excitation. Whenever two antennas are excited out of phase, a certain antenna pattern is produced. One easy way to obtain a  $180^\circ$  phase difference for exciting the antennas is with a half-wave r-f line. Figure 10-36 shows the use of a phase shifting line to provide the correct phase difference in the elements of the antenna array. The antenna on the left is excited by the r-f generator. The generator also applies the same signal to a half-wave line. Due to the phase shifting characteristics of the r-f line, the

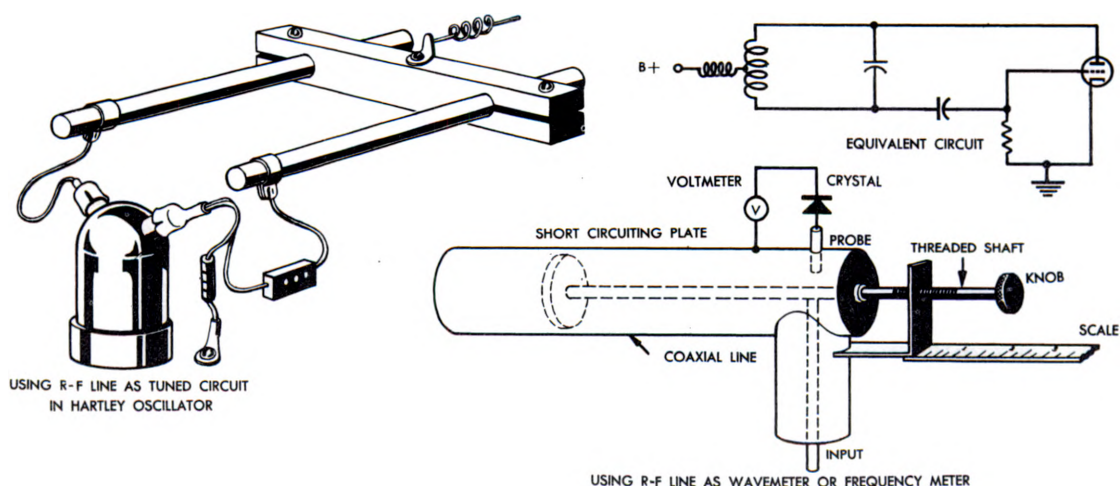


Figure 10-35. Using R-F Lines as Tuned Circuits

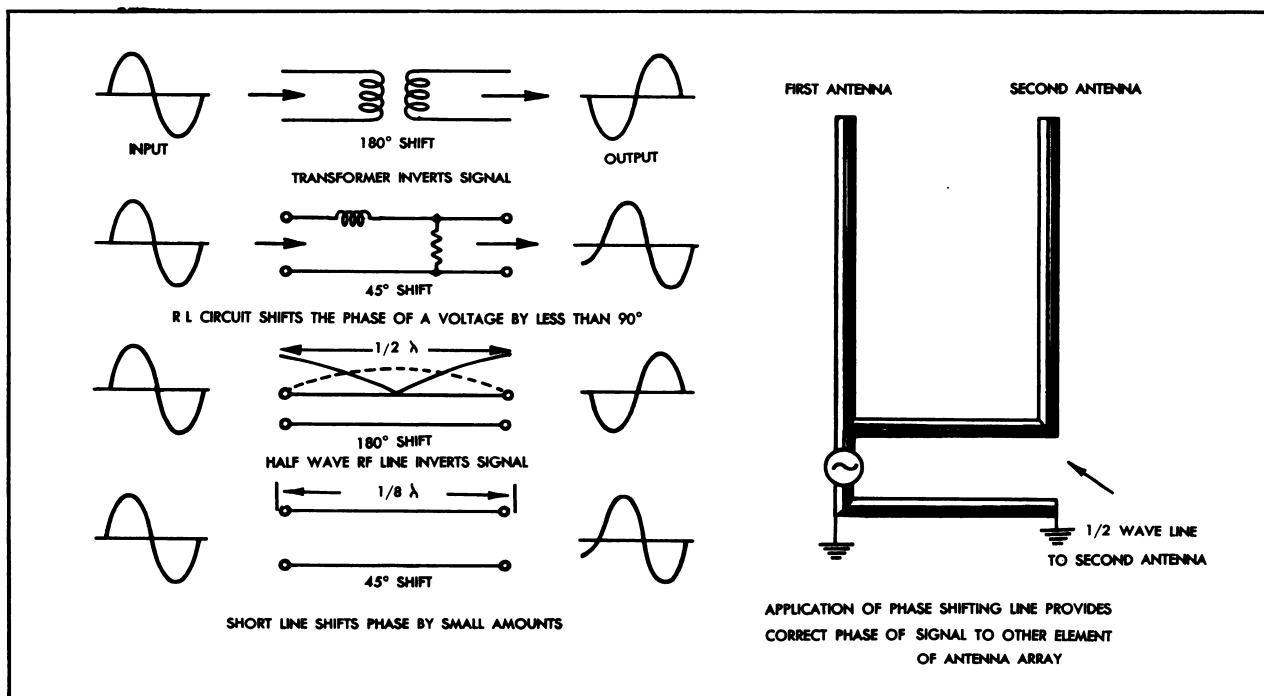


Figure 10-36. Shifting Phase with R-F Line Sections

antenna at the other end is fed a signal which differs  $180^\circ$  from that fed to the first antenna.

Other antenna directional characteristics can be obtained by varying the length of the r-f line. This variation of line length causes a change in phase which, in turn, gives different antenna radiation characteristics.

### ARTIFICIAL LINES

Although the delay characteristics and charging action of r-f lines make them useful in some radar applications, they cannot be used in others, chiefly because they would have to be long. An actual r-f line would have to be several hundred feet long for some uses in radar. Even if a line of such length could be coiled, it would occupy more space than is available in an aircraft. Thus, so far as airborne equipment is concerned, actual transmission lines are out of the question. For this reason artificial lines are used instead of actual lines.

Artificial lines possess all the characteristics of actual lines, but do not have the bulk and physical length of an actual line. They are constructed by first determining the capacitance and inductance

of the desired line, and then lumping these variables into equivalent inductors and capacitors. The result is an electrically equivalent, but physically different, line.

Figure 10-37A shows an actual two-wire line. Figure 10-37B shows the equivalent circuit for each section as it is when the section is split into two parts, and its characteristics measured. Since both sections have identical characteristics, the two can be joined together (Figure 10-37C). The final circuit, shown at D, lumps the inductive and capacitive characteristics of each section and simulates the artificial line. As you can see, the artificial line performs just like an actual line, but occupies much less space.

In any line time is required for a voltage change to travel the length of the line. This time characteristic makes it possible to slow down the transfer of a voltage change in its travel from one circuit to another.

The artificial line in Figure 10-38 provides a means of delaying a voltage pulse 10 microseconds. This line is constructed with an inductance of 0.5 millihenry and a capacity of 0.05 mf in each of the two sections. To compute the time for a voltage change to move through one sec-



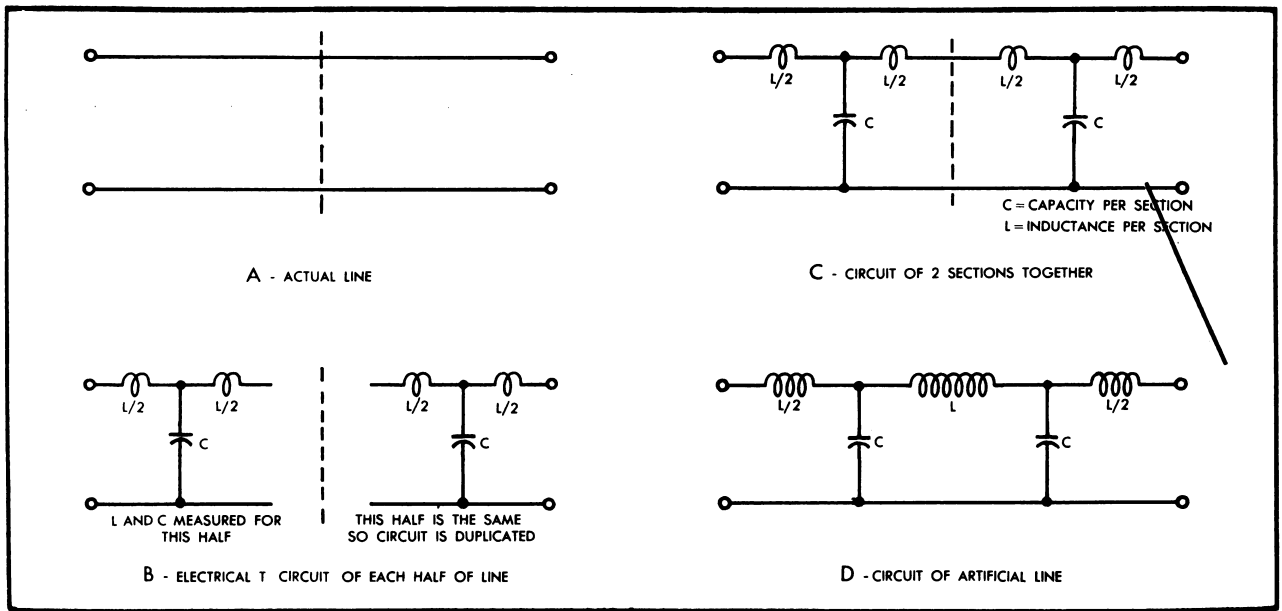


Figure 10-37. Design of the Artificial R-F Line

tion, use equation (9). The equation for time delay for the entire line is:

$$T_a = N\sqrt{LC} \quad (24)$$

where  $T_a$  is the time delay in seconds for the entire line,  $N$ , the number of sections, and  $L$  and  $C$ , the inductance and capacitance per section in henrys and farads respectively.

To calculate the time delay for the entire artificial line shown in Figure 10-38:

$$\begin{aligned} T_a &= N\sqrt{LC} = 2 \times \sqrt{0.5 \times 10^{-3} \times .05 \times 10^{-6}} \\ &= 2 \times \sqrt{5 \times 10^{-4} \times 5 \times 10^{-8}} \\ &= 2 \times \sqrt{25 \times 10^{-12}} \\ &= 2 \times 5 \times 10^{-6} \\ &= 10 \times 10^{-6} \text{ seconds} \end{aligned}$$

Time delay equals  $T_a = 10$  microseconds

In one radar set the time base is generated in such a way that it does not start until after the first radar echo returns to the receiver. For these signals to be displayed at their correct range mark, they are sent through an artificial r-f delay line which delays the signals just enough so that the very first returning signal does not arrive at the cathode-ray tube (CRT) until the time base has begun to form. Although a large number of video frequencies are present in the signal going through the line, all of them are delayed the same amount.

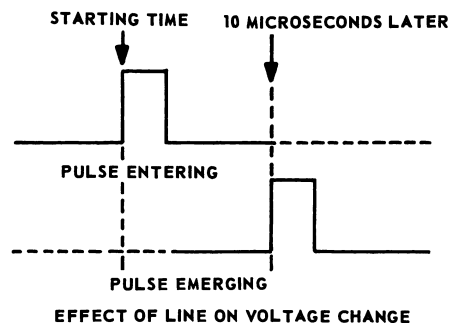
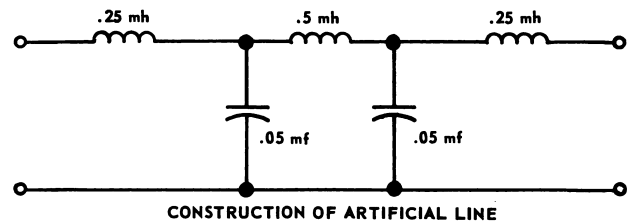


Figure 10-38. Artificial Delay Line

Another phenomenon of r-f lines is the manner in which the lines charge and discharge when a DC voltage is applied to them. When DC is applied to a line in which the source impedance is matched to the line impedance, half the source voltage appears across the line impedance at the time the battery is connected to the line. A

voltage change is produced across the line. Figure 10-39, which shows the oscilloscope picture at the sending end, illustrates this condition. This voltage change travels down the line, charging it as it goes. When the voltage change reaches the open end, it is reflected and starts back along the line.

All capacitors charge to half the battery voltage when the voltage change is going down, and to full battery voltage when the voltage rise is coming back. The oscilloscope picture in Figure 10-39, at the center of the line, illustrates this condition. This change in voltage is commonly called a *step* voltage. When all capacitors have been charged to a value equal to the battery voltage, current stops flowing from the battery. By observing the oscilloscope picture which illustrates the end of the line, you will note that the end of the line experiences a continuous rise of voltage which equals the applied voltage.

When this line is discharged into a resistance equal to its characteristic impedance, the result is a square wave that has a constant amplitude and a duration equal to twice the time for the pulse to travel the length of the line. The amplitude is equal to half the charge voltage because the line discharges through its own impedance and the load impedance ( $Z_R$ ), which are connected in series. This cuts the load voltage by

half and produces a drop of voltage that immediately travels down the line and back to the starting place as shown in Figure 10-40. As the end of the line is open, the reflection there is in phase. This produces a voltage across the load impedance ( $Z_R$ ) equal to half the charge voltage and which lasts for twice the time of travel on the line.

Usually an artificial line is constructed to generate this square wave. When the L and C for each section is known, the time duration of the square pulse,  $t_p$ , is equal to

$$t_p = 2N \sqrt{LC}$$

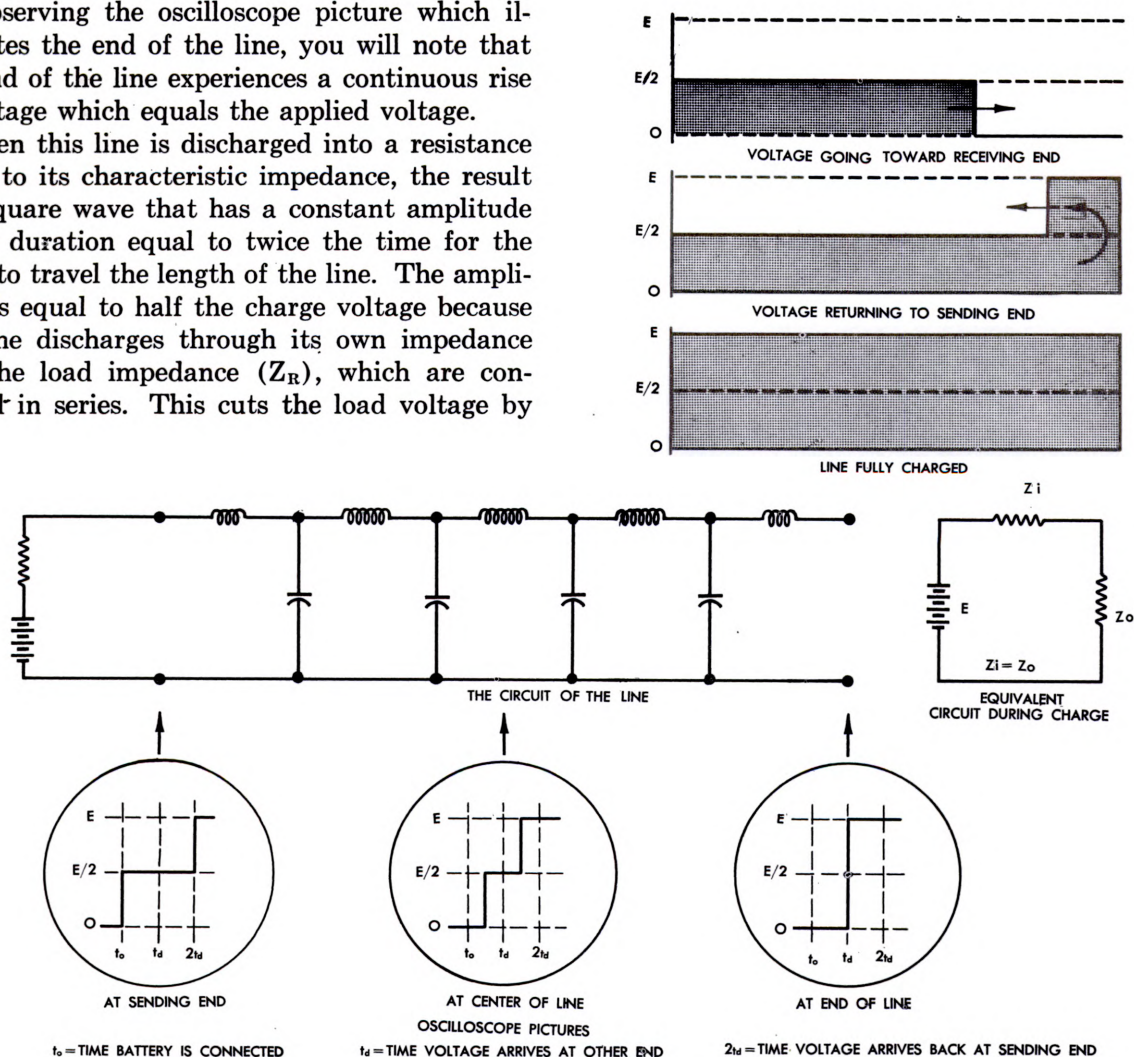


Figure 10-39. Mechanism of Charging Artificial Line



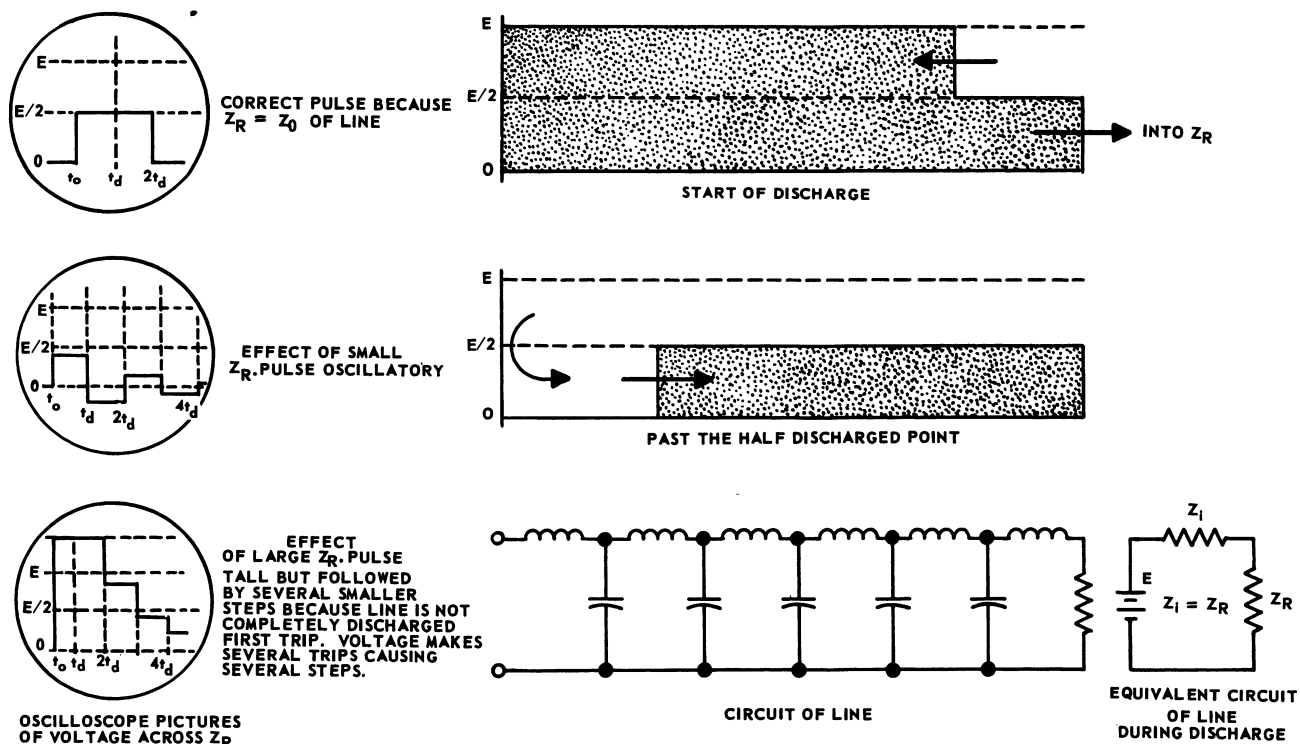


Figure 10-40. Discharging Artificial Line Through  $Z_R$

The following are the calculations for the time duration of the pulse generated in the artificial line in Figure 10-38:

$$L = 0.5 \text{ mh per section}$$

$$C = .05 \text{ mf per section}$$

$$N = 2 \text{ sections}$$

$$\begin{aligned} t_p &= 2N\sqrt{LC} = 2 \times 2 \times \sqrt{0.5 \times 10^{-3} \times .05 \times 10^{-6}} \\ &= 4 \times \sqrt{25 \times 10^{-12}} \\ &= 4 \times 5 \times 10^{-6} \\ &= 20 \times 10^{-6} \end{aligned}$$

The time duration is

$$t_p = 20 \text{ microseconds}$$

When an artificial line is not terminated in its characteristic impedance ( $Z_0$ ), several trips (reflections) are required for complete discharge. This means that discharge cannot be completed in a single round trip.

In Figure 10-40 there are three oscilloscope pictures that show the discharge action of an artificial line across  $Z_R$ . When  $Z_R$  is large, the pulse is tall. But it is followed by several smaller steps—called tails. These tails exist because the line did not completely discharge during the first trip and more trips are necessary for complete discharge. When  $Z_R$  is small, the change

that is reflected not only cancels the voltage left on the line, but also recharges the line to a smaller voltage in the opposite direction, producing an oscillatory output waveshape.

To calculate the  $Z_0$  of the line, use the formula

$$Z_0 = \sqrt{\frac{L}{C}}$$

The following are the calculations for finding the  $Z_0$  of the artificial delay line shown in Figure 10-38:

$$\begin{aligned} Z_0 &= \sqrt{\frac{L}{C}} = \sqrt{\frac{0.5 \times 10^{-3}}{5 \times 10^{-8}}} \\ &= \sqrt{\frac{5 \times 10^{-4}}{5 \times 10^{-8}}} \end{aligned}$$

The characteristic impedance is equal to

$$Z_0 = \sqrt{10^4} = 10^2 = 100 \text{ ohms}$$

An artificial pulse line is capable of producing a perfectly measured square wave pulse which has an amplitude equal to several thousand volts. As a matter of comparison, a vacuum tube multivibrator and amplifier circuit that could provide a similar output would weigh 100 pounds and occupy several cubic feet of space. On the other hand, a pulse-forming line of equal ca-

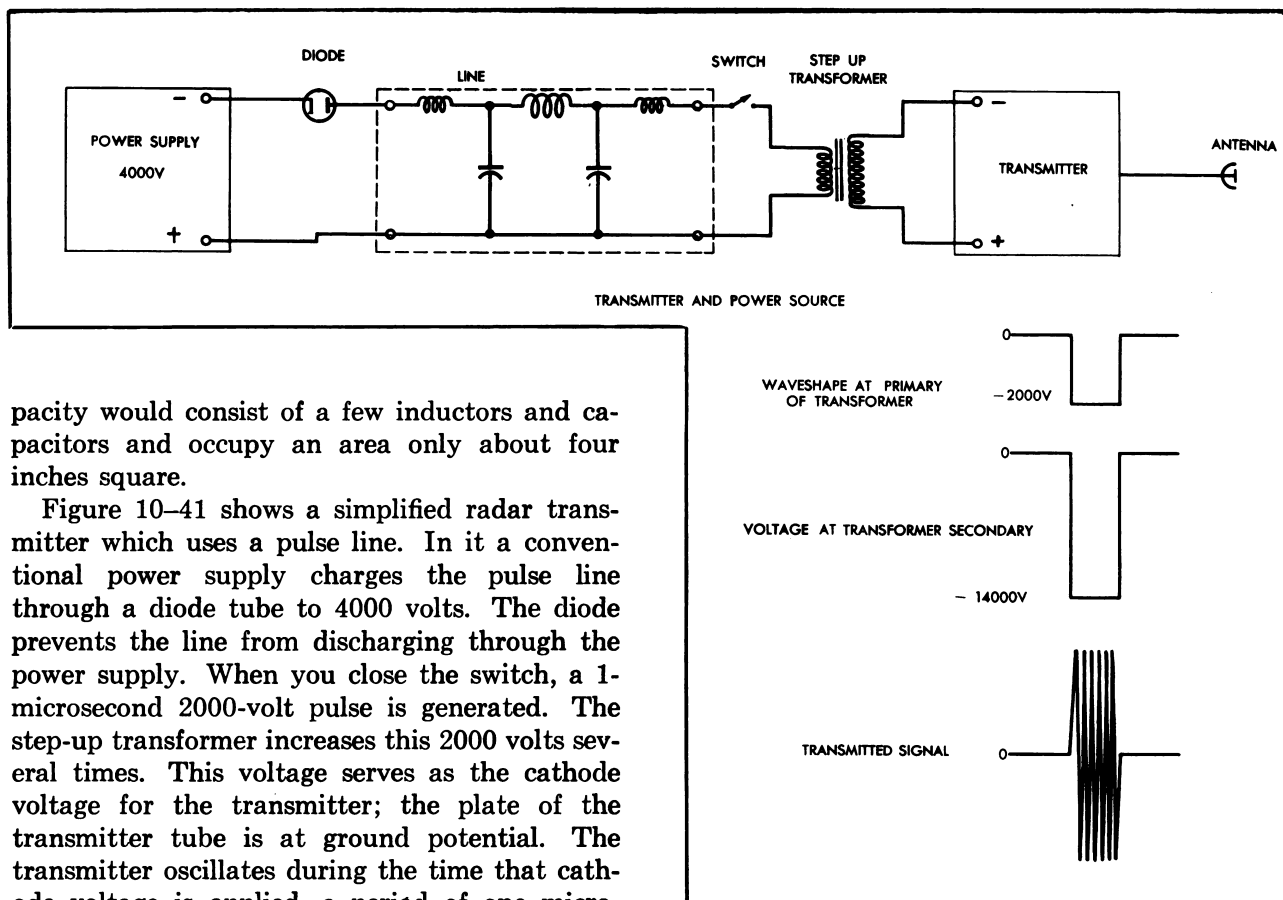


Figure 10-41. Simple Circuit Showing Pulse Line used in Radar Transmitter Circuit

capacity would consist of a few inductors and capacitors and occupy an area only about four inches square.

Figure 10-41 shows a simplified radar transmitter which uses a pulse line. In it a conventional power supply charges the pulse line through a diode tube to 4000 volts. The diode prevents the line from discharging through the power supply. When you close the switch, a 1-microsecond 2000-volt pulse is generated. The step-up transformer increases this 2000 volts several times. This voltage serves as the cathode voltage for the transmitter; the plate of the transmitter tube is at ground potential. The transmitter oscillates during the time that cathode voltage is applied, a period of one microsecond. Thus, a radio-frequency signal of one microsecond duration (or a radar pulse) is radiated by the antenna.

A pulse can also be formed by a short-circuited line. The operation of the line depends

upon the energy stored in the magnetic fields around the inductors in the line. The line is charged from a constant-current source and produces a voltage across the load ( $R_L$ ). Figure 10-42A shows a short-circuited line during charge. At the instant the constant-current source is connected to the line (T1 in Figure 10-42B), a current ( $I_0$ ) flows. If  $R_L$  is equal to the characteristic impedance of the line ( $Z_0$ ), the current is divided equally between the line and  $R_L$ . A wave of voltage whose amplitude is one-half  $I_0 Z_0$  travels down the line, and is reflected in *opposite* polarity at the short-circuited end. This reflected wave travels back to the input end.

The voltage pulse generated at the load during the charging period has an amplitude of  $\frac{1}{2} I_0 R_L$ . It lasts the length of time it takes the pulse to travel to the end of the line and back.

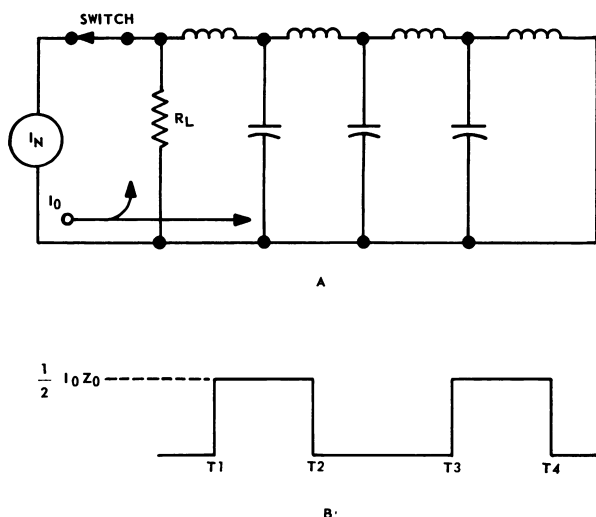


Figure 10-42. Pulse Formation with Shorted Line

This is the pulse shown from T1 to T2 in Figure 10-42B.

The voltage across  $R_L$  is now zero. (The line is effectively shorting the load.) A constant current  $I_0$  is flowing through the inductances of the line, and the line is fully charged with magnetic energy. The current flowing down the line during charge is reflected in phase, and the incident current and reflected current add, doubling the current on the line.

At time T3 the constant-current source is disconnected from the transmission line. A voltage pulse that is exactly similar to the charging voltage pulse is developed across  $R_L$  as the magnetic energy stored on the line is discharged. The duration of the pulse is the time it takes the current to travel down the line and back (T3 to T4). At T4 the line is discharged completely.

As the magnetic fields about the inductances collapse, half the current ( $\frac{1}{2} I_0$ ) flows through the load and half charges the capacitors. As the current tends to decrease through the inductors, the capacitors discharge, maintaining the current constant to the end of the pulse.

### SUPERSONIC DELAY LINES

The supersonic delay line has two general purposes: (1) to delay an electrical signal, such as a trigger or a range mark, and (2) to delay a pulse carrier without introducing distortion. In present-day radar systems comparison of video signals with each preceding signal is a necessity for efficient MTI operation. Recent advancement in the development of electronic equipment has indicated an increasing need of triggers whose timing cannot vary more than 0.5 microsecond.

The principal elements of a supersonic delay line are the transmission media and the electro-mechanical transducers. In the supersonic delay line electrical impulses are changed into sound impulses; after they have traversed a transmission medium, they are converted back into electrical signals. Since sound waves travel slower than electrical waves, delay is introduced. For frequencies in the megacycle range, quartz crystals are the most satisfactory transducers. Liquids (water and mercury) and solids (fused quartz) are used as the transmitting medium.

### Liquid Media

Figure 10-43 illustrates a typical mercury delay line. The signal is applied to a quartz crystal which is the electro-mechanical transducer. (A transducer is a device that converts electrical energy into mechanical energy, and vice versa.) The crystal undergoes changes in thickness due to the piezoelectric effect; electrical energy is converted into mechanical energy (vibrations). The vibrations of the crystal are communicated to the mercury as supersonic, or ultrasonic, waves. The waves travel down the tank with the mercury acting as a transmission medium.

On reaching the end of the tank, the waves are reflected by a reflecting plate. The reflectors consist of circular surfaces arranged on the two opposing ends of the tank. They are formed in the walls of the tank by grooves cut around these areas. If the reflecting surfaces make good contact with the mercury, they will tend to absorb the mechanical vibrations. For maximum reflection capabilities, a rough surface is used, with air trapped in the pockets of the rough surface area. This provides a mismatch to the mercury that causes almost perfect reflection. This process is repeated over several paths. The waves finally strike a crystal at the opposite end of the tank and produce electrical impulses. The piezoelectric effect here changes the mechanical energy to electrical energy.

In mounting the crystal care must be taken that it is supported and protected so that it will not bend or break. The crystal must be aligned accurately with the tube axis. Provision must be made to attach electrodes to the crystal surface so that there will be no interference with the crystal's sonic operation. Elimination of unwanted multiple echoes is a requirement for lines that must reproduce the pulse shape accurately. This is accomplished by reducing the time the signal is applied to the delay line so that the remaining wave can be damped out. The thickness vibration of the crystal is usually at the crystal's lowest resonant frequency. Some crystals are backed by steel on the sides not in contact with the mercury. In other crystal mountings the crystal is backed by mercury, which makes unnecessary any plating on that side. The large acoustic impedance of the mercury likewise loads the crystal and widens its bandpass.

In the mercury delay line shown in Figure 10-43, direct mercury-to-the-crystal contact is

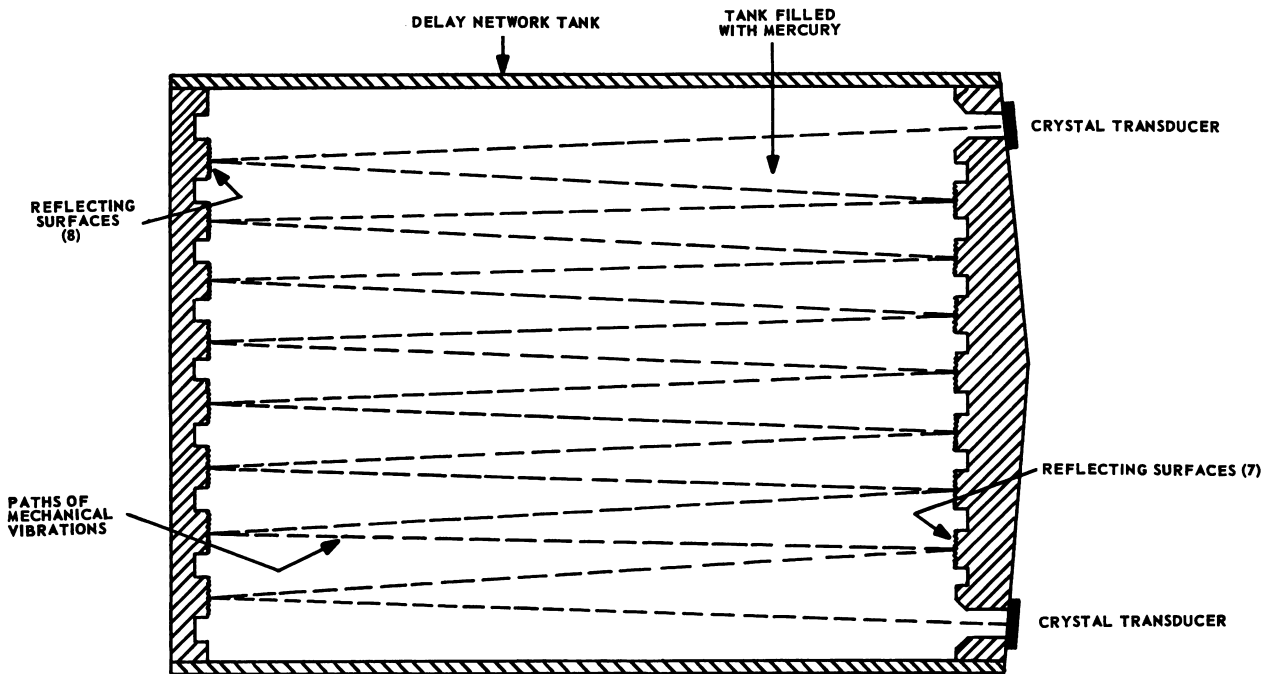


Figure 10-43. Delay Network Details

used at both the input and the output of the delay line. The mercury acts as the ground side of the electrical input and output circuits. Insulated plates on the opposite sides of the crystals provide the other electrical connections for crystal operation.

### Solid Media

Fused-quartz delay lines are mechanical delay circuits that depend upon the time it takes mechanical ultrasonic waves to pass through the fused quartz. Normally each quartz delay line is contained in an hermetically sealed shell. Figure 10-44 shows the *multiple symmetry* type of fused-quartz delay line. Transducer crystals with a resonant frequency near the frequency of the carrier signal convert electrical input waves to supersonic mechanical waves.

A polygon-shaped quartz block is used to provide surfaces from which the signals are reflected. The facets, or sides, of this polygon are precision ground at the necessary angles to obtain the desired signal reflection path. Figure 10-44 illustrates the theoretical polygon-shaped quartz crystal. The sides and angles of an actual fused-quartz delay line of 14 sides would vary. In Figure 10-44, the signal is fed in at side 2 and fed out at side 11. Some of the facets of the

multiple symmetry quartz delay line reflect the signal as many as three times. The illustrated figure shows only one of the possible paths which the signal may use.

Fused-quartz delay lines of the multiple symmetry type and mercury delay lines are used because of their ability to delay signals faithfully by much longer times than electrical delay lines. In the propagation of acoustical waves there is much less phase and amplitude distortion than in the electromagnetic waves. Such delay lines can also be constructed to withstand severe conditions of temperature and pressure and may be stored indefinitely without any harmful effects. Fused-quartz delay lines may be mounted in any position for operating.

One factor limiting operation of fused-quartz delay lines is that spurious secondary signals are produced in the line because of the signal traveling paths other than the desired path. These secondary signals are fed out of the line at times other than the desired delay time, and must be attenuated to a level at which they can be ignored. Another important factor concerning the use of the fused-quartz delay line is the amount of signal attenuation in the line. Some radar systems have a requirement of less than 60 db signal attenuation.

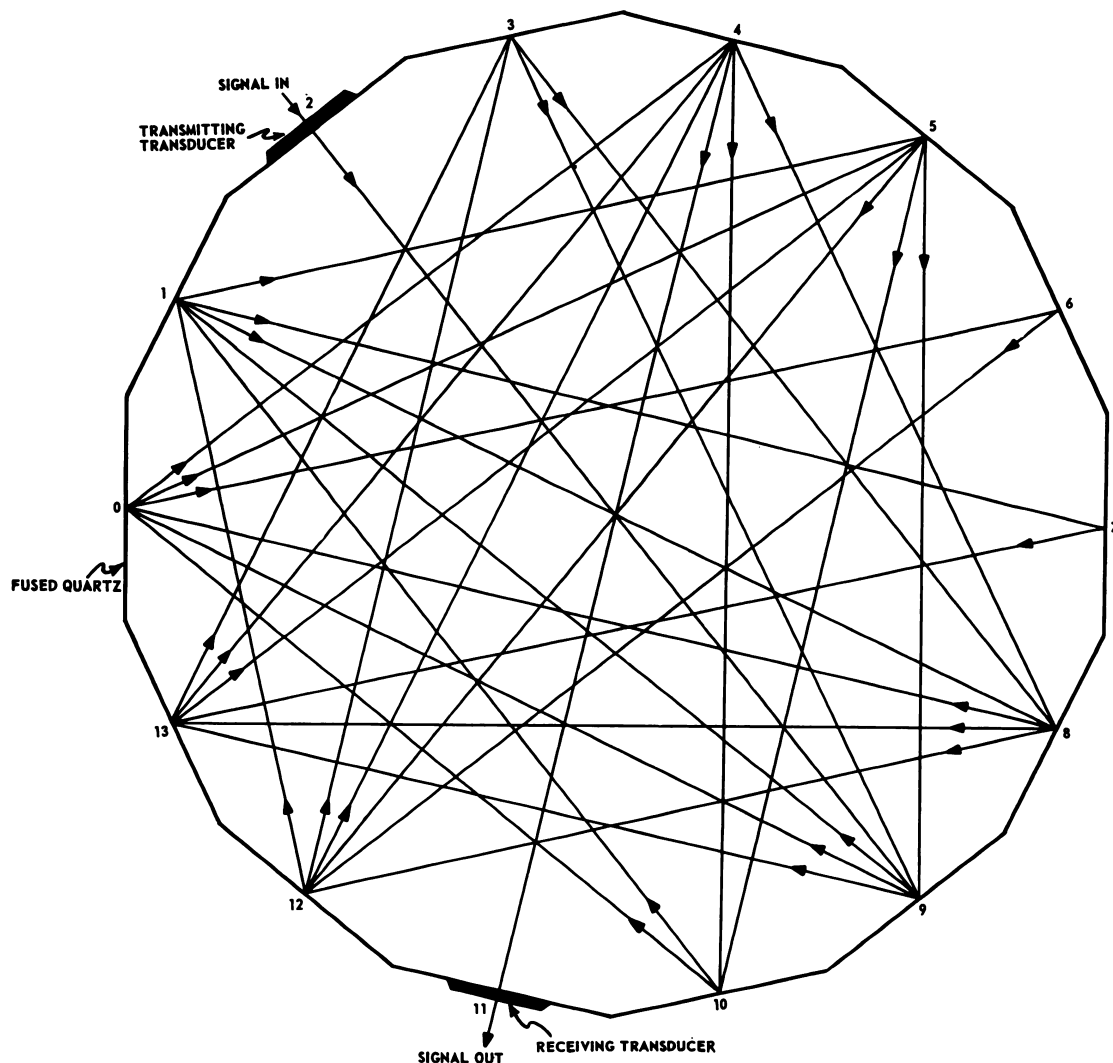


Figure 10-44. Quartz Delay Line

### Magnetostriction Delay Line

The basic material used in the construction of a magnetostriction delay line is a nickel ribbon or tube. Kinking should be avoided, as serious reflections occur at kinks or dents. The driving coil is approximately the same as the pickup coil, the pickup coil having more turns than the driver coil. A nonmetallic, nonmagnetic material is used as a coil mount, with the coil centered very accurately on the coil form. The output coil requires a polarizing magnetic field for proper operation. Undesired end reflections are prevented by some means of mechanical dampening, usually a rubber mount.

The main theory of ferromagnetism is based on the assumption that a ferromagnetic crystal

is composed of a large number of regions, each of which is spontaneously magnetized to an intensity that is a function of the temperature alone. When the net over-all magnetization is zero, these regions are aligned in equal numbers along the various directions of easy magnetization of the crystal. The effect of a magnetizing field is to alter this uniform distribution so that most of the regions lie in that direction of easy magnetization which most nearly coincides with the field.

The application of a magnetic field increases the stability of the atoms aligned with the field and decreases the stability of other orientations. As the field intensity increases, the atoms with unstable orientation shift suddenly. The cor-

responding sudden changes in the interatomic force of the face-centered cubes of nickel crystals set up elastic vibrations that act upon adjacent atoms and are thereby transmitted. The sum of a large number of these sudden local shifts in orientation results in the observable magnetostriction effect, which is used in present-day electronic equipment.

The effect may be described as the change in dimensions of a ferromagnetic material when subjected to a longitudinal magnetic field. The maximum variations occur in the length of the material with a smaller change occurring in the volume of the material.

In Figure 10-45,  $V_1$  is a twin triode whose grids are biased beyond cutoff. A positive pulse drives both grids of  $V_1$  positive, causing a short pulse of current to flow in each plate circuit. Each plate circuit of the twin triode contains 3 coils, with a set of relay contacts in parallel with each coil. The driver coils are wound around the magnetostriction line of nickel tubing. The relays are remotely controlled. If any relay contact is opened, the plate current pulse flows through the associated coil and creates a magnetic field along the longitudinal axis of the nickel tubing. This field sets up a shock wave

in the tubing because of the magnetostriction effect.

A mechanical stress is felt along the tubing in the vicinity of the driver coils. The shock wave travels from its source toward the end of the line with a velocity of 0.187 inch per microsecond. When this shock wave reaches the end of the line, most of the energy is absorbed in the rubber tubing that is clamped around the end of the line. The driver coils may be spaced so that the time required for the shock wave to travel from one coil to another meets the time requirements of the equipment used.

A permanent horseshoe magnet is mounted near the tubing and provides a magnetic field in the vicinity of the pickup coil. The shock wave, which actually compresses the nickel tubing, travels down the tubing and shifts the lines of flux created by the permanent magnet. As the lines of flux cut the pickup coil, a voltage is induced. Because of the short duration of the shock wave and the direction of the magnetic field, a short negative pulse is applied to the grid of  $V_2$ . The magnetostriction delay line has been developed to a satisfactory state as a flexible device for delaying pulses, mixing pulses and generating codes.

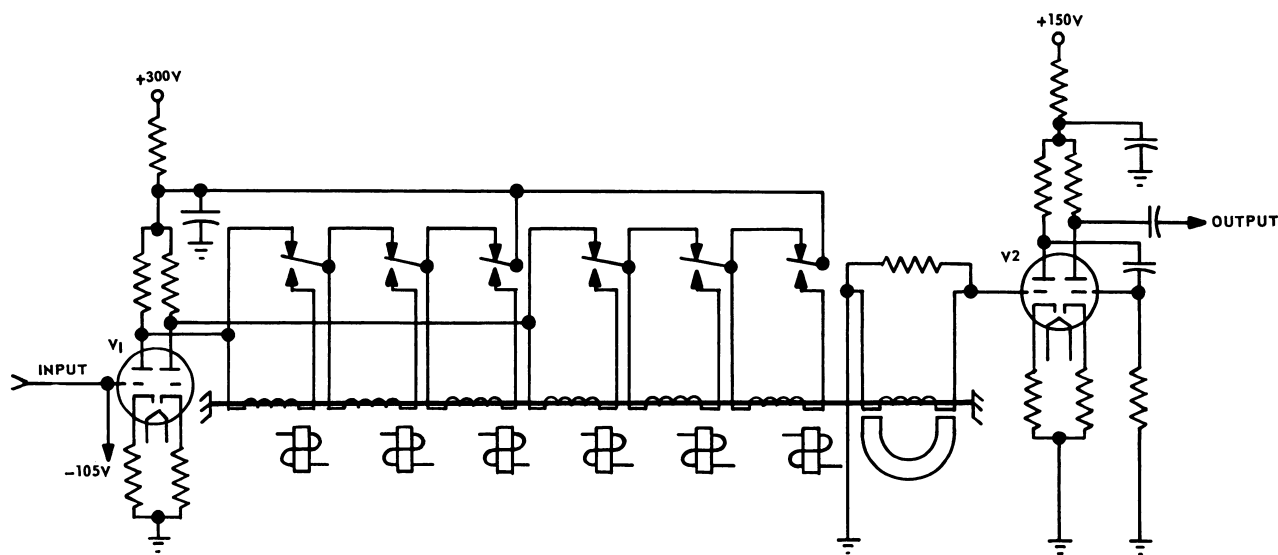


Figure 10-45. Magnetostriction Delay Line



# Waveguides, Cavity Resonators, and Duplexers

The Air Force operates some electronic equipment at frequencies above 1000 mc. The very short wavelengths of these frequencies give them the name of *microwaves*. The transmission lines that are used at these frequencies are different both physically and electrically from the ones previously discussed.

Since conductors radiate the energy they carry, and since radiation increases with an increase in frequency, special transmission lines are used at microwave frequencies. If a microwave signal were applied to a two-wire line, the radiation would increase to such an extent that practically no energy would reach the output end.

Even though the inner conductor of a coaxial line radiates energy, the coaxial line can be used for microwave transmission. All of the energy is confined between the outer and inner conductors. Removing the center conductor and transmitting the wave through the remaining conductor results in a transmission line called a cylindrical waveguide. When a hollow rectangular conductor is used, the transmission line is called a rectangular waveguide.

Another device that is used at microwave frequencies is a cavity resonator. The purpose of this chapter is to acquaint you with these devices. It discusses their theory, operation, and use.

## GUIDED AND UNGUIDED ELECTROMAGNETIC WAVES

In general, there are two methods for transferring electrical energy—one is by current flow in wires; the other is by movement of elec-

tromagnetic fields in space. Electrical energy can be transferred as current flow in a number of types of transmission lines, for example, two-wire lines and coaxial lines; in space it moves as electromagnetic fields whenever a radio or radar antenna radiates energy.

Electromagnetic fields that move in space are confined largely to the area between the earth and the ionosphere as you can see in Figure 11-1A. (Chapter 15 explains the ionosphere.) While the change in the dielectric constant of the ionosphere is sufficient to reflect most electromagnetic fields that strike it, extremely high frequency radiation that strikes it almost perpendicularly is not reflected. An electromagnetic field that does not start out in a direction parallel to the earth and the ionosphere, follows a zig-zag path between these two areas and may or may not be reflected back to the earth.

Although the transfer of energy by electromagnetic fields and by currents in wires may seem to be unrelated phenomena, actually the trend on the part of electronic scientists is to look even on two-wire lines as elements that guide electromagnetic fields from one place to another. The currents in the wires are merely considered incidental to the action and the result of the moving fields.

A two-wire line is a poor guide for transferring electromagnetic fields because it does not confine the fields in a direction perpendicular to the plane that contains the wires as shown in Figure 11-1B. This results in some energy escaping in the form of radiation. Electromagnetic



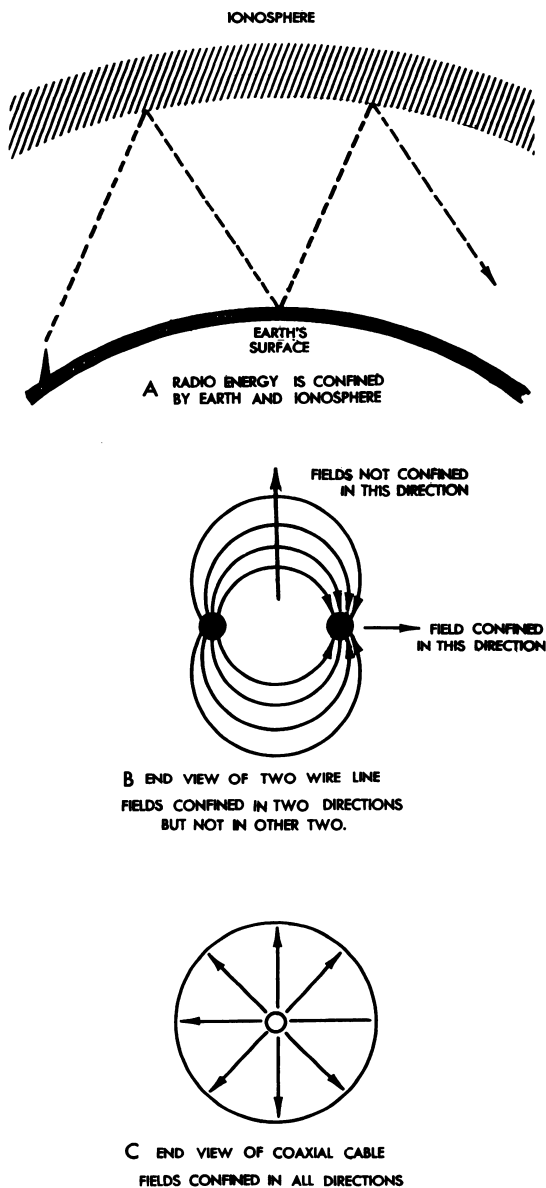


Figure 11-1. Guiding Waves

fields may be completely confined in this direction when one conductor is extended around the other to form a coaxial cable as shown in Figure 11-1C. In a coaxial cable, energy transfer is said to take place by electromagnetic fields, rather than by current flow. However, this method is not too efficient at high radio frequencies, since skin effect limits the current-carrying area of a conductor to a thin layer at its surface.

Another disadvantage is that the ability of the fields to form is limited by the amount of

current flow associated with them. When the resistance in a conductor is increased, current flow in it is reduced, the reduction depending on the amount of increase. This decreases the magnitude of the fields. On inspecting the cross section of the coaxial cable, you can see that the surface area of the inner conductor is much less than the surface area of the outer conductor. This causes the inner conductor to retard the current considerably more than the outer conductor and results in a reduction in the efficiency of energy transfer. If you could remove the center conductor and retain the fields, energy might be transferred with less loss.

Electromagnetic fields can transfer energy in a line that does not have a center conductor provided the configuration of the fields is changed to compensate for the missing conductor. The area remaining, which might be considered a pipe, is called a *waveguide*. A waveguide does not necessarily have to be circular in cross section. Practical waveguides, for example, are also square, rectangular, or elliptical in construction.

Metallic walls are not necessary to guide electromagnetic fields in a waveguide, for the fields will be reflected whenever they encounter any kind of a substance which has a different dielectric constant from the substance in which they are traveling. For example, fields can be made to travel through a ceramic rod with little loss of energy. When they encounter the air at the surface of the rod, they are reflected back into the rod.

As you previously learned, the three types of losses in r-f lines are copper losses, dielectric losses, and radiation losses. Briefly, these losses are described as follows: Copper loss is an  $I^2R$  loss. It becomes appreciable whenever skin effect reduces the conducting area of the lines. Dielectric losses are losses caused by the heating of the insulation between the conductors. Radiation losses are losses due to energy escaping from r-f lines in the form of radiation. Let us consider waveguides from the point of view of these losses.

Copper losses are small in waveguides. Since a two-wire line consists of a pair of small conductors, the surface area of each is likewise small. Although the surface area of the outer conductor of a coaxial cable is large, the inner conductor is small, and it produces considerable copper losses.

On the other hand, a waveguide has a large surface area as it does not have a center conductor. Therefore, whenever current flows, the copper losses in the waveguide are less than those in other types of lines.

Dielectric loss of energy is eliminated in a waveguide, since there is no center conductor requiring a solid dielectric support. One of the major reasons for signal attenuation in a coaxial line is the leakage in the dielectric used to support the inner conductor. Because the dielectric material has some resistance, a part of the energy transmitted down the line is absorbed by the insulator.

The frequency of the transmitted energy is the principal factor governing dielectric loss. As the frequency is increased, the amount of electromagnetic energy absorbed by the dielectric also increases. The dielectric loss is therefore much higher at microwave frequencies than at the lower radio frequencies.

Radiation losses are less in a waveguide than in a two-wire line. In a waveguide, fields are contained wholly within the guide itself just as in a coaxial line. Therefore, only a negligible amount of energy is radiated.

The power handling capacity of a waveguide is greater than that of a coaxial line having an equal size. Power is a function of  $E^2/Z_0$ , where  $E$  is the maximum voltage in the traveling wave and  $Z_0$  is the characteristic impedance of the line.  $E$  is limited by the distance between the conductors. In the coaxial line illustrated in Figure 11-2, this distance is  $S_1$ . In the waveguide this distance, which is  $S_2$ , is much greater than  $S_1$ . Therefore, the waveguide is able to handle greater power before the voltage exceeds the breakdown potential of the insulation.

A waveguide is simpler to construct than a coaxial line. This is because in the waveguide the center conductor is eliminated completely.

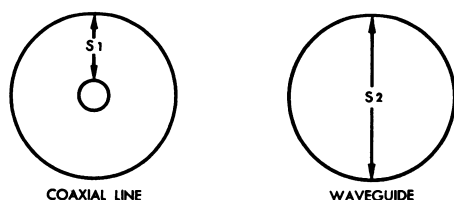


Figure 11-2. Comparison of Spacing in Coaxial Line and Waveguide

The physical stamina of a waveguide is greater than that of a coaxial line. This is due to the fact that, unlike a coaxial line, it has no center conductor or insulators which can be displaced or broken.

In view of these advantages you may wonder why waveguides are not used exclusively for transferring energy. There are, however, two disadvantages which make waveguides impractical to use at any but extremely high frequencies.

In the first place, the cross-sectional dimensions of a waveguide must be in the order of a half-wavelength for it to contain electromagnetic fields properly. A waveguide used at one megacycle, for example, would be about 700 feet wide. At lower radar frequencies, 200 mc for example, this waveguide would have to be about four feet wide, while at higher radar frequencies, such as 10,000 mc, it need be only one inch wide. Therefore, dimensions that waveguides require make them impractical at any frequency lower than about 3000 mc.

In the second place, if the dimension of the guide is a half-wavelength or less, energy will not be propagated through the waveguide. The reason for this is that for any given waveguide there is a cutoff frequency, below which it does not function as a power transfer device. This limits the frequency range of any system using waveguides.

The installation of waveguides presents problems that are different from those encountered with two-wire or coaxial lines. Waveguide installations are often referred to as *electrical plumbing*, because the exact positioning and rigidity required calls for installation techniques well known to plumbers, but less known to electronic or microwave technicians.

Special couplings are required to fasten the lengths of the waveguide together. At the lower radio frequencies adequate transmission characteristics are assured by soldering the conductors together. At higher microwave frequencies precautions to prevent the leakage of any energy must be taken. An ideal situation is to have the waveguide one continuous section between transmitting and receiving points. But the normal situation has the waveguide made in fairly short sections that must be connected together. Figure 11-3 illustrates some of the waveguide sections manufactured for specific applications. Special flexible waveguide sections have been

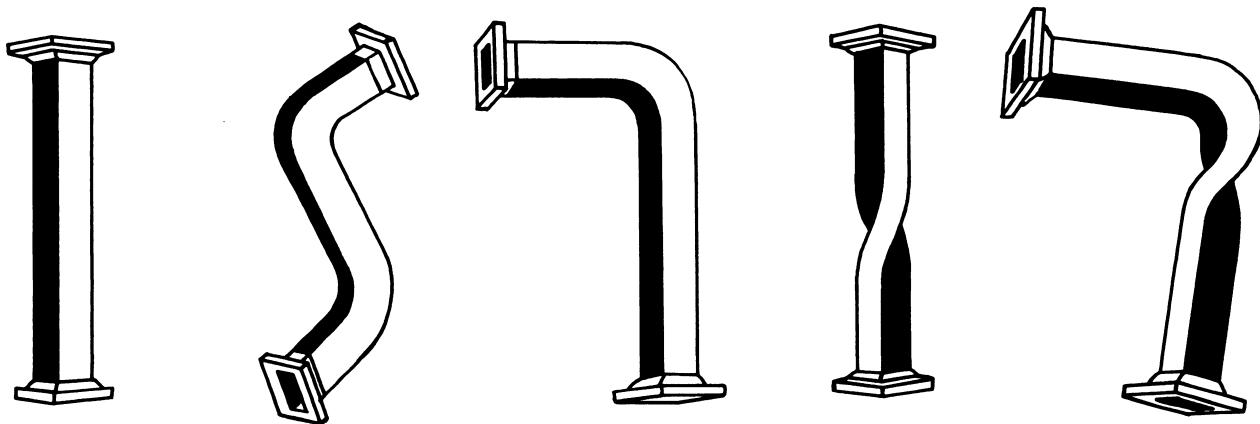


Figure 11-3. Waveguide Sections

developed to be used where the standard, or fixed, curved sections cannot be used.

It is easily understood why each new installation must be carefully planned beforehand. After the initial planning the various sections of waveguide required can be ordered and the whole line can be assembled. Compare this type of installation to that encountered when flexible coaxial cable, which is easily bent around corners or passed through wiring troughs, is used. You can understand that waveguide installations are more complex and more expensive.

Besides installation costs, the price of the waveguide is high. The waveguide must meet rigid mechanical specifications to prevent discontinuities along the transmission path which might introduce energy losses. During the manufacture of the waveguide, special machining is performed to minimize such discontinuities. This special machining also adds to the initial cost. To reduce the skin-effect losses sometimes the inside of the waveguide is silver-plated or even gold-plated. This also results in increased cost.

### WAVEGUIDE THEORY

An exact mathematical analysis of the way in which fields exist in a waveguide is quite complicated. But it is possible to obtain an understanding of many of the properties of waveguide propagation by using the following simple analogy, which shows both how the fields are able to exist in a waveguide and how you can handle them.

To understand the action of a waveguide, assume that a waveguide has the form of a two-wire line. In this condition there must be some means of supporting the two wires. Furthermore, the support must be a nonconductor, so that no power will be lost by radiation leakage. An efficient way for both insulating and supporting the two-wire line is shown in Figure 11-4A. This line is spaced, insulated, and supported by porcelain stand-off insulators. At communication frequencies, the absorption of power by the dielectric material (insulators) causes them to look like a low resistance and capacitance.

The equivalent electrical circuit at higher frequencies is shown in Figure 11-4B. For frequencies of 3000 mc and up, a better insulator than nonconducting porcelain insulators must be used. A superior high-frequency insulator for this purpose is a quarter-wave section of r-f line, called a *metallic* insulator, which was discussed in the preceding chapter. Such an insulator is shown in Figure 11-4C. As there are no dielectric losses in a quarter-wave section of r-f line, the impedance at the open end (the junction of the two-wire line) is very high.

A metallic insulator can be placed anywhere along a two-wire line. Figure 11-5A shows several on each side of a two-wire line. A point to note in this line is that the supports are a quarter-wave at only one frequency. This limits the high efficiency of the two-wire line to one frequency only.

The use of several insulators results in the improved conductivity of a two-wire line when

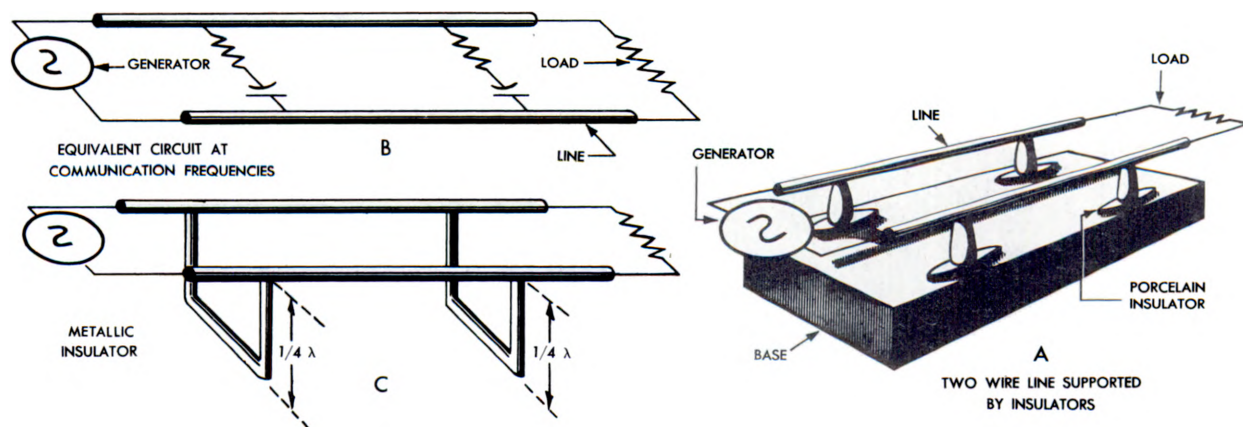


Figure 11-4. Insulating the Two-Wire Line

the sections are connected together. This connection is made between the two adjacent insulators through a switch, as shown in Figure 11-5B. When the switch is open, both quarter-wave sections are excited by the main line. In this condition, there will be standing waves on the quarter-wave sections.

When the switch is connected to the same place on each section, the relative phase relationship of the voltages at the connection will be the same for each section. In this condition the No. 1 section will be excited first by the generator. When the switch is closed, the No. 2 section will be partly excited by the No. 1 section through the switch connection. In this manner, less energy from the main line is required to excite the No. 2 section. The parallel paths shown cause less resistance to exist along a given length of line, and energy is transferred with less copper loss.

When more and more quarter-wave sections are added to the line until each section makes contact with the next, the result is a rectangular box in which the line is at the center, Figure 11-5C. The line itself is actually part of the wall of the box. The rectangular box thus formed is a waveguide.

#### EFFECT OF DIFFERENT FREQUENCIES ON A WAVEGUIDE

Previously it was stated that a quarter-wave section is limited in operation to a certain frequency. However, when a solid wall of insulators

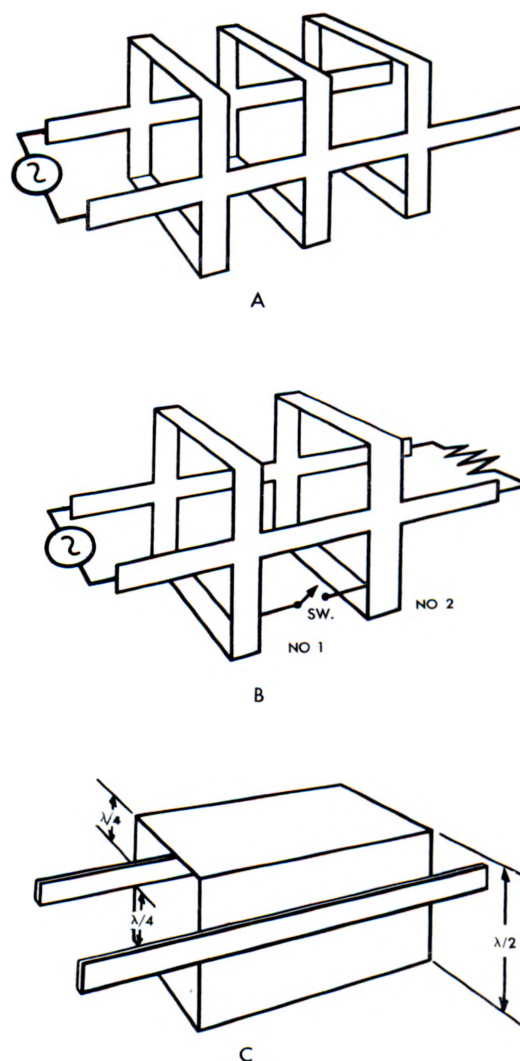


Figure 11-5. Development of Waveguide by Adding Quarter-Wave Sections



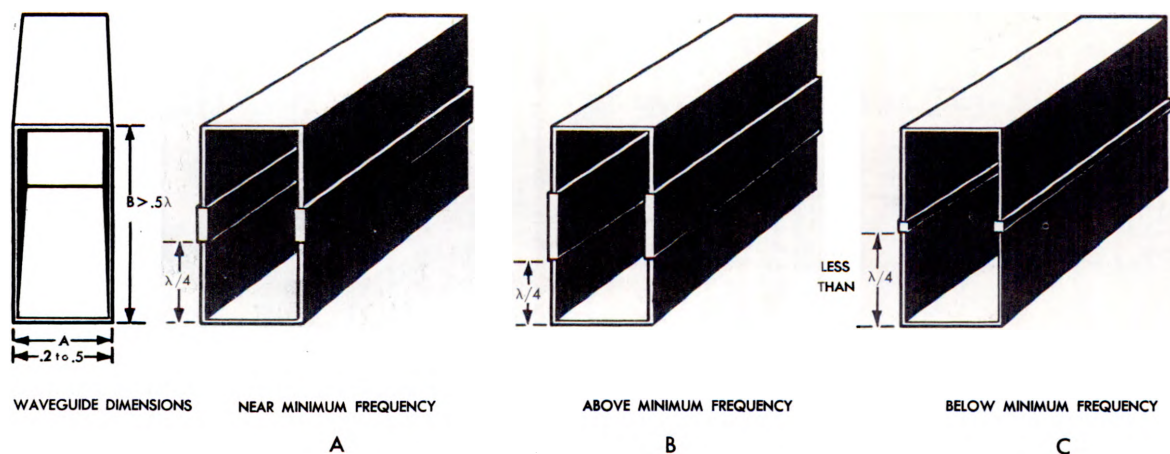


Figure 11-6. Effect of Different Frequencies in the Waveguide

is added, the section will operate at other frequencies. The waveguide shown in Figure 11-6A is the one just discussed. When the frequency being transferred by the guide is made higher, the quarter-wave section must be shorter. These shorter wavelengths are easily accommodated if you assume this two-wire line is made up of a wide bar or strip in each wall of the guide, as shown in Figure 11-6B. The shorter distance remaining is the shorter quarter-wave section.

Thus, the wide bar shown is theoretically well insulated at any frequency higher than the one which creates the near minimum frequency at A. In reality there is a practical upper frequency limit at which this analogy is applicable. For example, when the bar is a half-wavelength across, the waveguide will be 4 quarter-wavelengths across and may act as though there were two bars instead of one, as shown later in the chapter.

The next consideration is a frequency which is lower than the original frequency. Lowering the frequency in a given waveguide will lengthen the sections and narrow the bar. Beyond some lower frequency, this bar does not exist, because the quarter-wave sections meet one another. At a still lower frequency, the sections become less than a quarter-wave as in Figure 11-6C. A section less than a quarter-wave is inductive. Thus, the impedance across the place where the conducting bars belong is not a high resistance, but an inductance. The inductive reactance will dissipate the energy in the line very swiftly through high currents that flow back and forth in one

place, as though the inductance were taking energy during one half-cycle and then returning it to the line during the next half-cycle.

Therefore, we find that in a waveguide there is a low-frequency limit, or *cutoff* frequency, below which the waveguide cannot transfer energy. The width of the waveguide at the cutoff frequency is equal to a half-wavelength, since at this frequency the two quarter-wave sections touch and add to one another. Mathematically, the cutoff wavelength is expressed by the equation,

$$\lambda c = 2B$$

where  $\lambda c$  is the cutoff wavelength, and  $B$  is the long dimension of the rectangular cross section.

As cutoff occurs when the width of a waveguide is below a half-wavelength, most waveguides are made 0.7 wavelength in the wide dimension to give a margin between the actual size and the size for cutoff. The other dimension is the distance between the conductors, and similarly, as in the two-wire lines, is governed by the voltage breakdown potential of the dielectric, which is usually air. Widths of 0.2- to 0.5-wavelength are common.

## ELECTROMAGNETIC FIELDS IN A WAVEGUIDE

A good working knowledge of the fields present in a waveguide is necessary if you want to use them intelligently. Energy in a waveguide is transferred by the electromagnetic fields, while currents and voltages merely aid in forming these

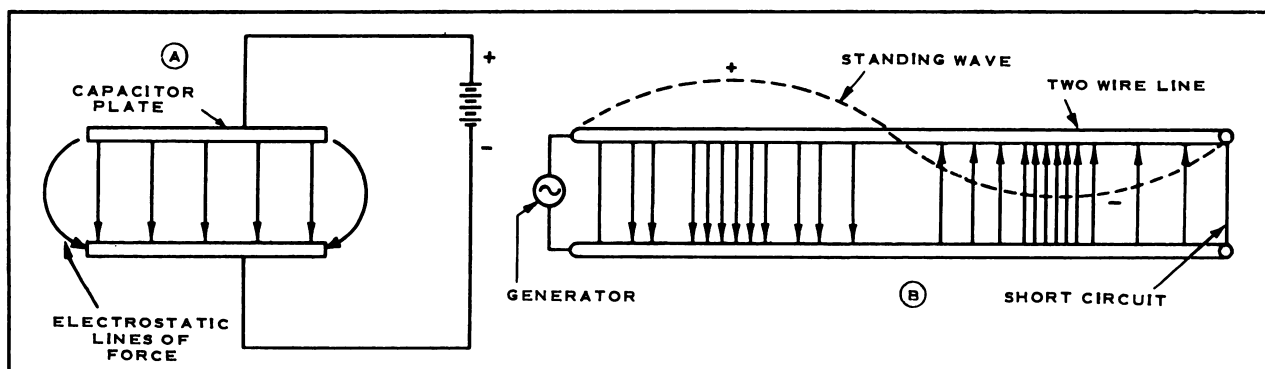


Figure 11-7. Electric Field Between Capacitor Plates and a Full-Wave Section of Two-Wire Line

fields. You should know when a good current path is required and where the voltages will be high. As energy is normally introduced into and removed from the waveguide by the fields, you should know where the fields will exist. In any waveguide two fields — the electromagnetic and the electrostatic field — are always present.

### The Electric Field

The existence of an electrostatic (electric) field indicates that there is a difference in the number of electrons between two points. An electric field consists of a stress in the dielectric field and is represented in diagrams by arrows. The simplest form of electrostatic field is the field that forms between the two parallel plates of a capacitor as is shown in Figure 11-7A. When the top plate of the capacitor is made positive by a battery, electrons move from the top plate and deposit themselves on the bottom plate. This immediately sets up a stress in the dielectric between the plates. This stress is represented by arrows, pointing from the more positive voltage point to the more negative voltage point. The amount of stress sometimes is indicated by the length of the arrow. A long arrow represents more stress than a short arrow.

In representing electrostatic fields, the number of arrows indicates the strength of the field. In the case of the capacitor in Figure 11-7A, note that the arrows are evenly spaced across the area between the two plates. As the voltage across the plates is the same at all points, the electrostatic lines between the plates are evenly distributed. This set of lines represents an electrostatic field. This field is usually called the

*electric field* and is abbreviated the *E-field*. The lines of stress are called the *E-lines*.

Notice in Figure 11-7B the two-wire transmission line which has an instantaneous standing wave of voltage applied to it. This line is equal to one wavelength. At the same time, part of it is positive, while another part is negative. The instantaneous electrostatic field (E-field) is the same at the negative and the positive points, but the arrows representing each field point in opposite directions. The voltage along the line varies sinusoidally. Therefore, the density of the E-lines varies sinusoidally.

An easy way to show the development of the E-field in a waveguide is from a two-wire line which has quarter-wave insulators. Figure 11-8 shows the two-wire line previously discussed with several double quarter-wave insulators, or half-wave frames used as insulators. The E-field on the main line is the same as that in the transmission line illustrated in Figure 11-7. The half-wave frames located at points of high voltage (strong E-field) will have a strong E-field across them. The half-wave frames located at points of minimum voltage will have no E-field on them.

Frame No. 1 in Figure 11-8 is an example of an insulator which has a strong E-field across it. Each frame is shown separately below the main line for a clearer view. Frame No. 2 is at a zero voltage point, so it will have no field on it. Frame No. 3 also has a strong field, but its polarity is reversed. Frame No. 4 has a weaker field on it due to its being at a lower voltage point on the main line. The picture shown is a build-up to the three dimensional aspect of the full E-field in a waveguide.

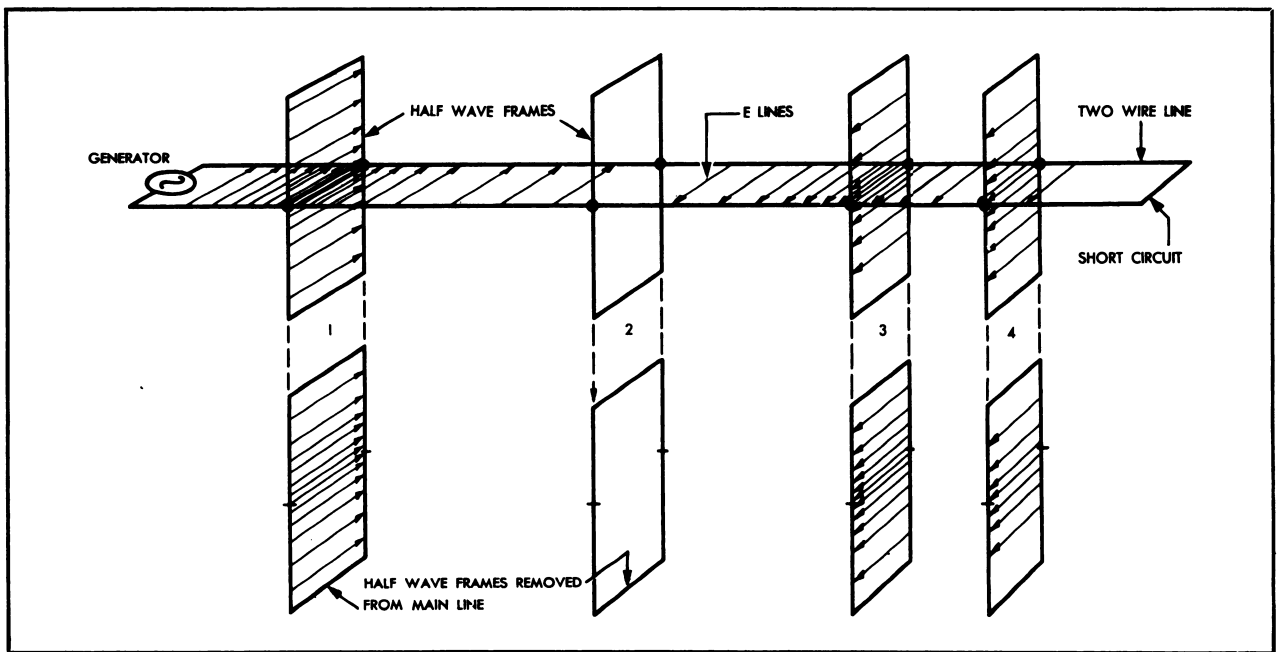


Figure 11-8. Magnitude of Fields on Half-Wave Frames Vary with Strength of Field on Main Line

The illustrations in Figure 11-9 show the E-field in an actual waveguide. This is the field that results when an infinite number of quarter-wave sections are connected to the line to form a rectangular box. The E-field is strong at one-quarter and three-quarter distances from the

shorted end, but becomes weaker at the sine rate toward the upper and lower walls and toward the ends and center. Again the phenomenon of wavelength is present, as shown in Figure 11-9A. You should realize, of course, that this is an instantaneous picture taken at the time the

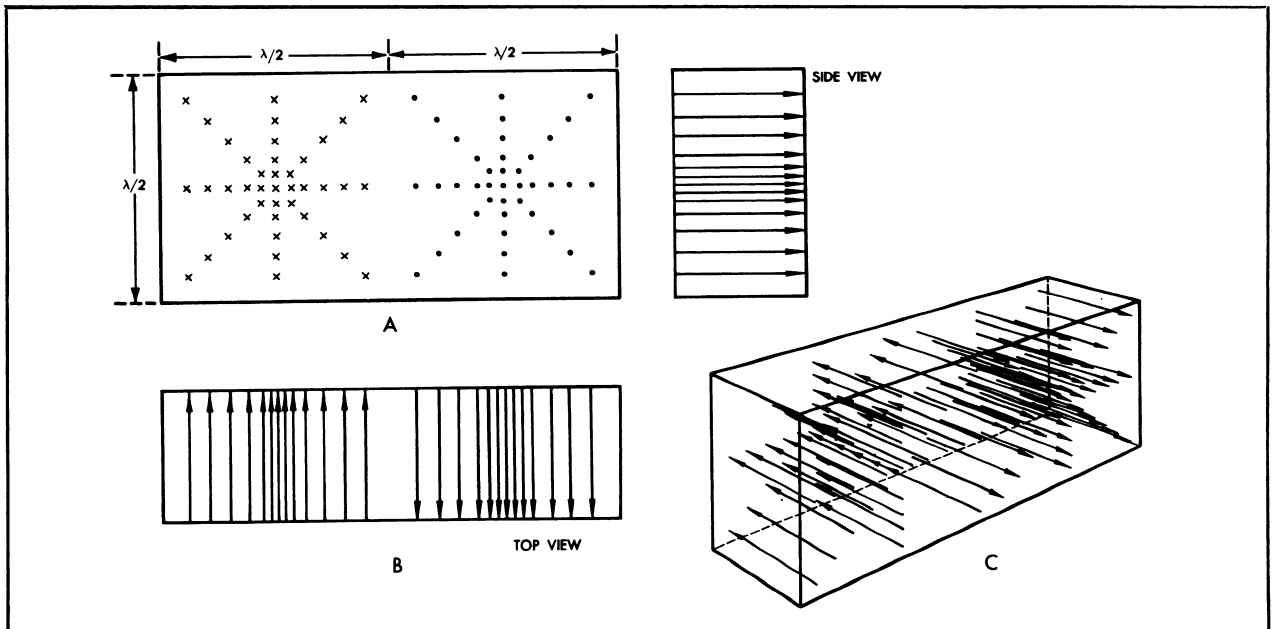


Figure 11-9. E-Field in Actual Waveguide



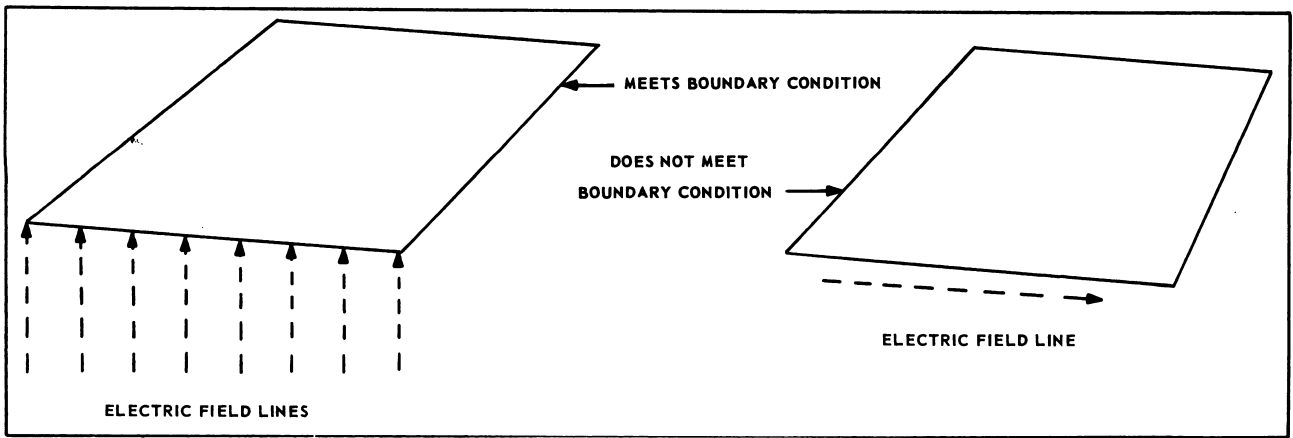


Figure 11-10. Boundary Condition for an Electric Field

standing wave of voltage is at its peak. At other times, the voltage and E-field vary from zero to the peak value, reversing direction every alternation of the applied voltage.

### Boundary Conditions

Certain boundary conditions must exist for propagation to occur in a waveguide. The propagation of energy down the waveguide is similar to that of electromagnetic waves transmitted in free space. The exception is that waves in the waveguide are confined to the boundaries of the metallic guide. When a waveguide is being used for transmission, certain *boundary conditions* have to be satisfied; otherwise, transmission will not continue.

There are two boundary conditions that have to be satisfied. The first is that an electric field

must be perpendicular to the conductor to exist at the surface of a conductor. The converse of this statement is also true. Assuming a perfect conductor, if an electric field has a component that is parallel to the conductor, an electric field cannot exist at the surface of a conductor. Figure 11-10 illustrates the first boundary condition. The second is that a varying magnetic field must exist in closed loops parallel to the conductors and perpendicular to the electric field.

Normally, if a transmitting system satisfies one boundary condition, it also satisfies the other condition. To check for one condition is all that is normally required to determine whether a wave can exist at the surface of a conductor.

Figure 11-11 illustrates a plane electromagnetic wave that is radiated and is confined between two parallel plate conductors. To satisfy

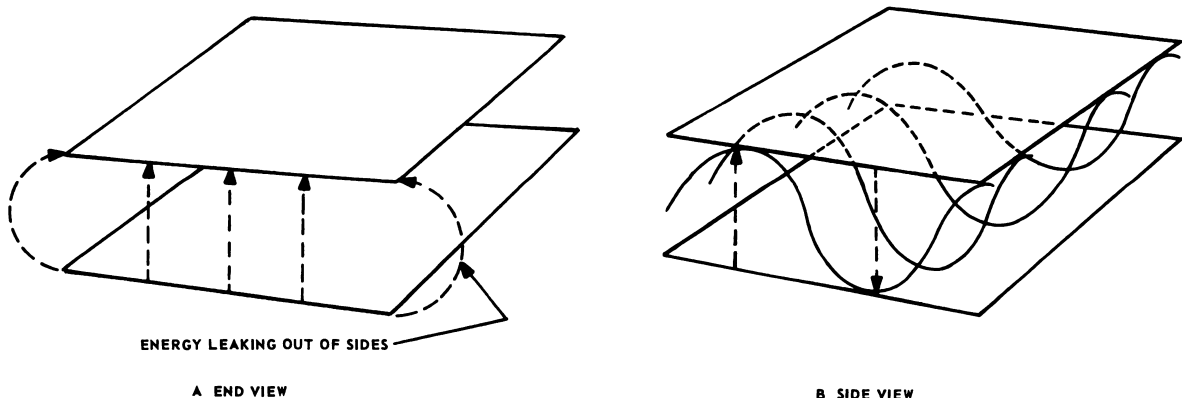


Figure 11-11. Electric Field between Plates

one boundary condition, the electric field is oriented so that the electric field (E) lines are perpendicular to the conductors as shown in Figure 11-11A.

Figure 11-11B shows an E-field that is varying sinusoidally along the distance between the plates. A varying E-field produces a varying magnetic field, which must exist in closed loops and must be parallel to the conductors, the second boundary condition. Figure 11-12 shows the closed loops of the magnetic field that is between the conductors at a given instant of time.

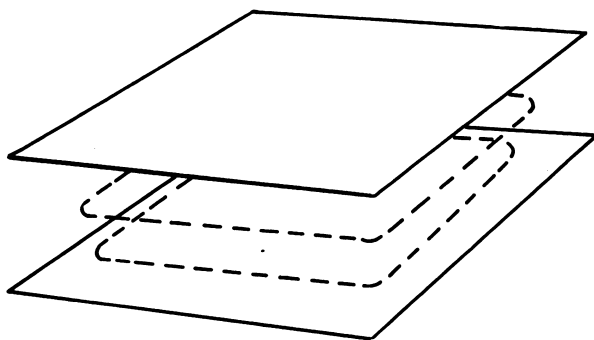


Figure 11-12. Magnetic Loops between Plates

Figure 11-11A also illustrates how some of the transmitted energy might leak or escape through the open sides. To prevent this energy

loss, side walls are attached to the parallel plates. A rectangular waveguide is thus formed. However, the plane wave shown cannot exist in a rectangular waveguide because the plane wave has an electric field that is parallel to the surface of the side walls.

### The Magnetic Field

The second field that must always exist in a waveguide is the magnetic field. The magnetic lines of force which make up the magnetic field are caused by the *movement* of electrons in the conducting material. All the tiny magnetic forces exerted by the individual moving electrons add together and form a large force around the conductor. The presence of the force is shown by closed loops around the single wire in Figure 11-13A. The line forming the loop is a magnetic line of force or an *H-line*. The H-line must be a continuous closed loop in order to exist.

All the lines associated with current are collectively called a magnetic field or an *H-field*. The strength of the H-field varies directly with the current. Each H-line has a certain direction. You can determine this direction by the left-hand rule. The strength of the H-field is indicated by the number of H-lines in a given area.

Although H-lines encircle a single straight wire, they behave differently when the wire is formed into a coil as is shown in Figure 11-13B. In a coil the individual H-lines tend to form

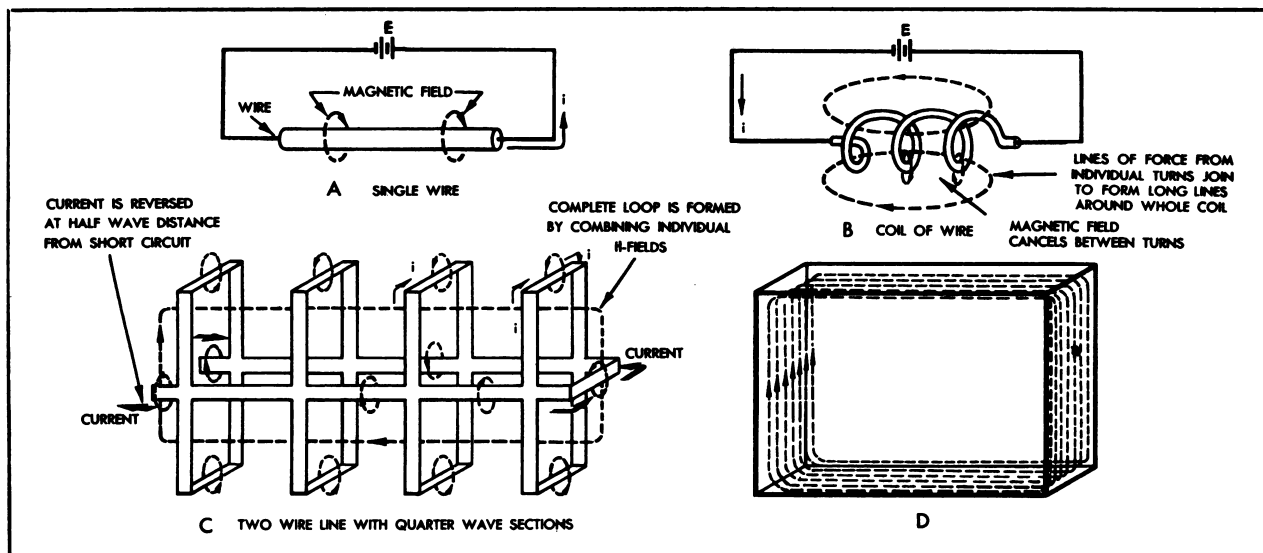


Figure 11-13. Development of Magnetic or H-Field in the Waveguide

around each turn of wire, but in doing so take opposite directions between adjacent turns. This causes cancellations and results in zero field strength between the turns. But inside and outside the coil, the directions are the same for each H-field. Therefore, the fields here join and form a continuous H-line around the entire coil as shown.

Similar action also takes place in a waveguide. In Figure 11-13C a two-wire line with quarter-wave sections is shown. Currents flow in the main line and in the quarter-wave sections. The current direction produces the individual H-lines around each conductor as shown. When a large number of sections exist, the fields cancel between the sections, but their directions are the same inside and outside the waveguide.

At half-wave intervals on the main line, current will flow in opposite directions. This produces H-line loops having opposite directions. In Figure 11-13C, current at the left end is opposite to the current at the right end. The individual loops on the main line are opposite in direction. All around the framework they join so that the long loop shown in Figure 11-13D is formed. Outside of the waveguide the individual loops cannot join to form a continuous loop. Thus, there is no magnetic field outside of a waveguide.

Figure 11-14 shows a conventional presentation of the magnetic field in a waveguide three half-wavelengths long. Note that the field is strongest at the edges of the waveguide. This is where the current is the highest. The current is lowest at the center of each set of loops because there the standing wave of current is zero at all times. In Figure 11-14 the picture shown

represents an instantaneous condition. During the peak of the other half-cycle of AC input, all field directions are reversed. An instantaneous picture of this condition would show the fields reversed at half-wave intervals, since the current in the long line is reversed over half-wave distances.

A second boundary condition necessary for electromagnetic fields to transfer power is satisfied by the configuration of the magnetic field. This condition, previously stated, requires that at the surface of the waveguide there be no perpendicular component of the magnetic field. Since all the H-lines are parallel to the surface, this condition is satisfied.

Electric and magnetic fields exist simultaneously in the waveguide. In fact, the H-field causes a current which, in turn, causes a voltage difference. This produces an E-field and it, in turn, causes a current. This current creates an H-field, and so on. One field is dependent on the other, as energy is continually transferred from one field to the other.

Figure 11-15A illustrates both fields in the waveguide. Since this picture is rather complicated, the presence and direction of the field is usually indicated in more simple diagrams, such as those shown in Figures 11-15B, C, and D. In these diagrams the number of E-lines in a given area indicates the strength of the electrostatic field, while the number of H-lines in any given cross section indicates the strength of the magnetic field in that area.

The field configuration shown in Figure 11-15 represents only one of the many ways in which fields are able to exist in a waveguide. Such a

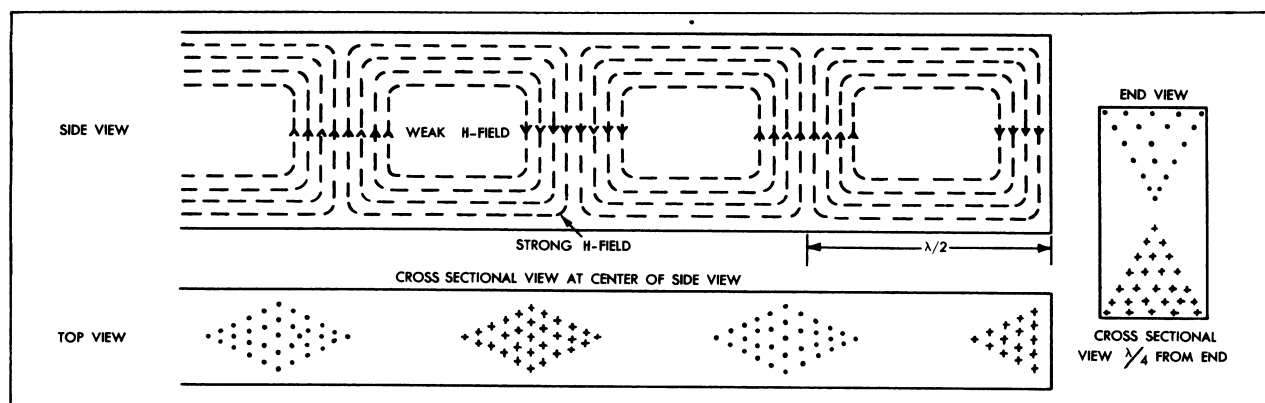


Figure 11-14. Magnetic Field in Waveguide Three Half-Wavelengths Long

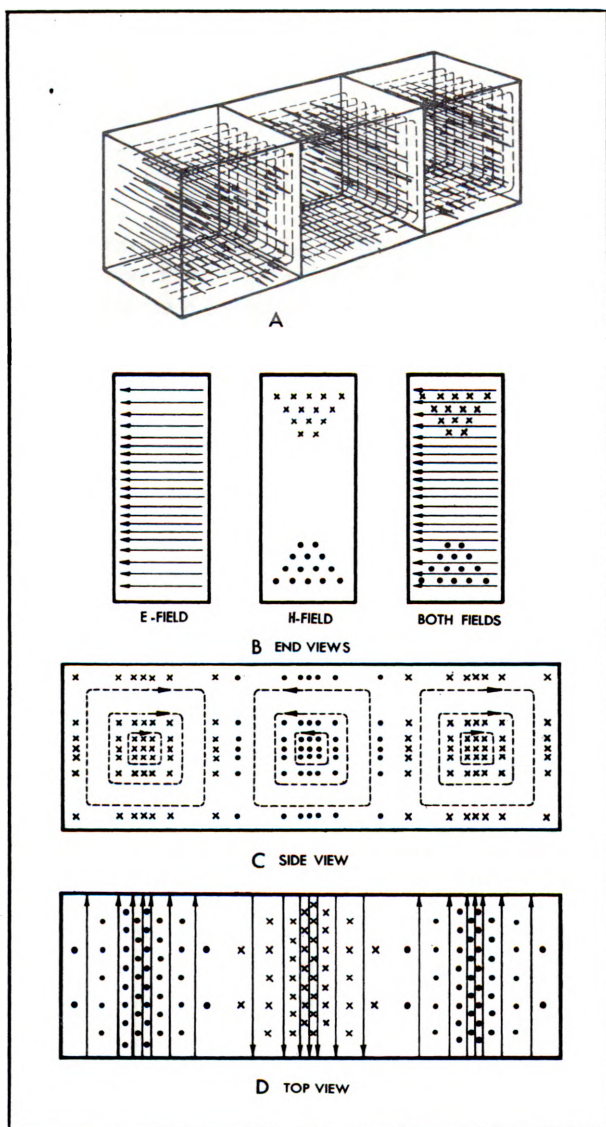


Figure 11-15. Conventional Picture of Both Fields in Waveguide

field configuration is called a *mode* of operation. In the case of the rectangular waveguide illustrated, the configuration is known as the *dominant* mode, since it is the easiest one to produce. Other higher modes — that is, different field configurations — may occur accidentally or may be caused deliberately in a waveguide.

An example of another field configuration is developed in Figure 11-16A. If the size of this waveguide is doubled over that of the waveguide shown in Figure 11-15, the cross section will be a full wave rather than a half-wave. The two-wire conductor can be assumed to be a quarter-

wave down from the top (or a quarter-wave up from the bottom). The remaining distance to the bottom is three-quarters of a wavelength. A three-quarter-wave section has the same high impedance input as the quarter-wave section. Thus, the two-wire line is properly insulated and will transfer energy. The field configuration will show a full wave across the wide dimension, as you can see in Figure 11-16B.

This field configuration can be applied to a circular waveguide. The two conductors shown in Figure 11-16C are assumed to be part of the waveguide wall. The remaining part of the wall forms the quarter-wave sections. The quarter-wave section insulates the two conductors. This makes it possible to transfer energy with minimum losses. The resulting field configuration shown at D is the dominant mode for a waveguide with a circular cross section.

#### ANALOGY OF WAVEGUIDE ACTION BY ELECTRIC WAVES

A somewhat different analogy involving waveguide action deals with the field rather than with current and voltages. This analogy is somewhat more exact than the previous explanation, which dealt with voltages and current in two-wire lines, for power flow in this case is assumed to be in the fields rather than in the conductors.

In a waveguide the fields are the same as those radiated into space by an antenna. Figure 11-17 shows a small portion of a field that is radiated into space from an antenna. In it the electrostatic lines of force, or E-lines, are parallel to the antenna, and the electromagnetic lines of force, or H-lines, are perpendicular to the antenna. They move away from the antenna at the speed of light. At each half-cycle the polarity is reversed. Therefore, at half-wave intervals, the fields are in opposite directions (or polarity). Although only a small part of the total field is shown, the E-lines and H-lines form huge closed loops after they leave the antenna.

As previously mentioned, the energy which moves through a waveguide and the energy which is radiated by an antenna are both the same form of electromagnetic radiation. Nevertheless, the field configuration shown in Figure 11-18 cannot exist in a waveguide because it does not satisfy the required boundary conditions. First, there cannot be any E-lines tangent to the sur-



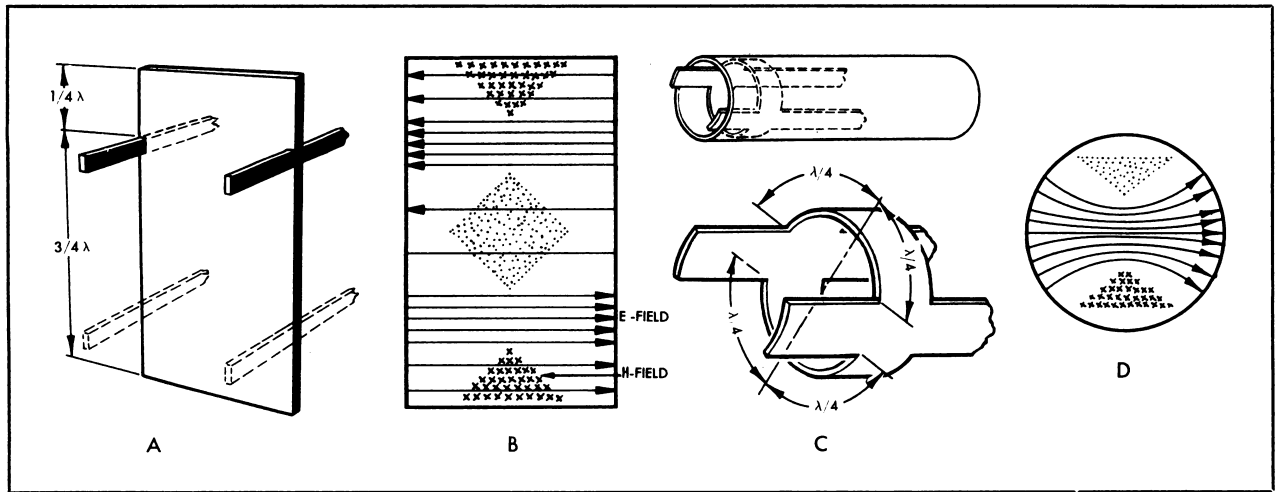


Figure 11-16. Another Field Configuration in the Rectangular Waveguide

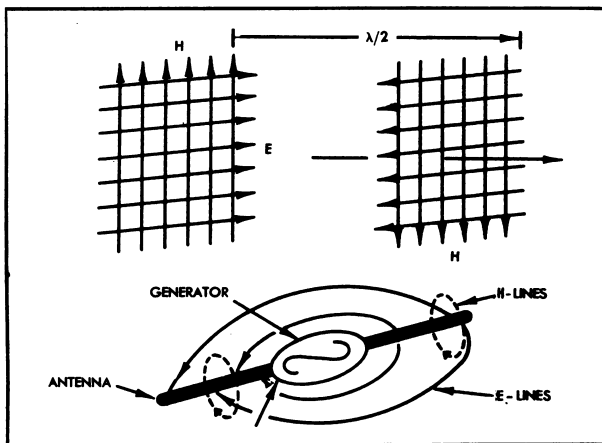


Figure 11-17. Small Portion of Field Radiated into Space by an Antenna

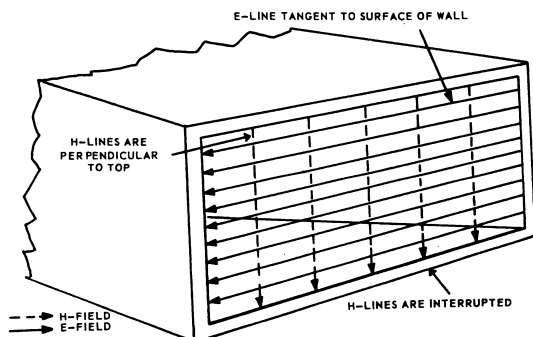


Figure 11-18. Fields in a Waveguide Must Satisfy Boundary Conditions to be Radiated

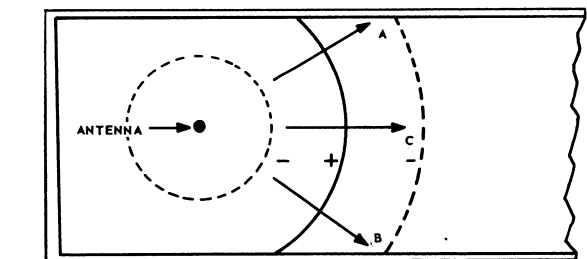


Figure 11-19. How Radiation Fields are Made to Fit a Pipe

face of the walls. Since the E-lines are evenly distributed across the area, some will be across the top and bottom wall. This causes the voltage to short out and this, in turn, causes the E-lines to be removed. Other E-lines which are pushed up to the wall by the repulsion between E-lines likewise short out. This shorting out is accumulative and eventually removes the entire E-field.

The second boundary condition which must be satisfied is that there must be no component of the magnetic field perpendicular to the wall. Note again in Figure 11-18 that the H-lines are parallel to the side walls, which is correct — but are perpendicular to the bottom, which cannot be; so H-lines of this type cannot exist in the waveguide. Furthermore, an H-line cannot exist unless it is a closed loop.

When a small antenna is placed in the waveguide and excited at a radio frequency, both positive and negative half-cycles are radiated as shown in Figure 11-19. The wavefront produced

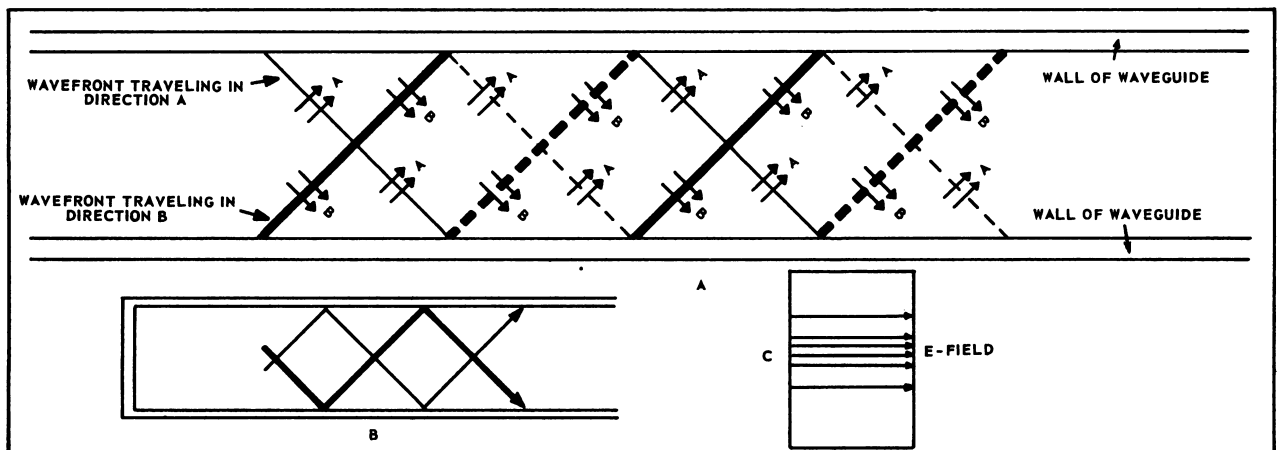


Figure 11-20. Paths of Wavefronts in Waveguide

is like an expanding circle. The part which travels in the direction of arrow C goes straight down the waveguide and is quickly attenuated, as previously described. However, the part of the wavefront which travels in the direction of arrow A is reflected from the wall. The wall is a short circuit and causes the wavefront to be reflected in reverse phase. Meanwhile, the wavefront that travels in direction B is reflected from the other wall and proceeds in opposite phase. Thus, the radiation fields are contained in the waveguide as illustrated in Figure 11-20B.

#### Paths of Wavefronts in a Waveguide

Look through the top wall of the waveguide in Figure 11-20A, the light solid and broken lines represent the wavefront going in direction A in Figure 11-19. The heavy lines and dashes represent the wavefront going in direction B. Note that all parts of the wavefront A are traveling upward at an angle across the guide. Wavefront B is traveling at the same angle but downward.

When the wave travels in this fashion in a waveguide, propagation is possible. In understanding what happens, note that the positive wavefront (represented by solid lines) occurs simultaneously throughout the center of the guide. These fronts add and cause a maximum voltage to occur at the center. (The E-field is shown maximum at the center in Figure 11-20C.) The negative wavefront adds in the same manner as the positive wavefront. When the negative wavefront meets the positive wavefront at the

walls, the two wavefronts cancel each other, making the total voltage equal to zero. This verifies the E-field condition shown in Figure 11-20C.

#### Crossing Angle

The angle at which a wavefront crosses a waveguide is a function of the wavelength and the cross-sectional dimension of the waveguide. At some intermediate frequency the reflection is as shown in Figure 11-21B. But as the frequency increases, the angle of incidence becomes less and the signal travels farther before it reaches the other side (see Figure 11-21A). At lower frequencies the wavefront crosses the guide at more nearly right angles to the walls. At some frequency the angle will be  $90^\circ$ . At this point the wave travels back and forth across the guide until the energy is dissipated by the resistance of the walls of the guide. At this frequency the distance from side to side is a half-wavelength for the waveguide. At the cutoff frequency the attenuation is a linear function of length and is very high.

The velocity of propagation of a wave along a two-wire line is less than its velocity in air. The same is true in a waveguide. Movement of a wave along a two-wire line is slower than its movement in air because of the retarding effect of the DC resistance of the conductors and the conductance of the insulation. In the waveguide the lower velocity is due to the way the field travels. As shown in Figure 11-21C, the path of a wavefront at a relatively low frequency is along the

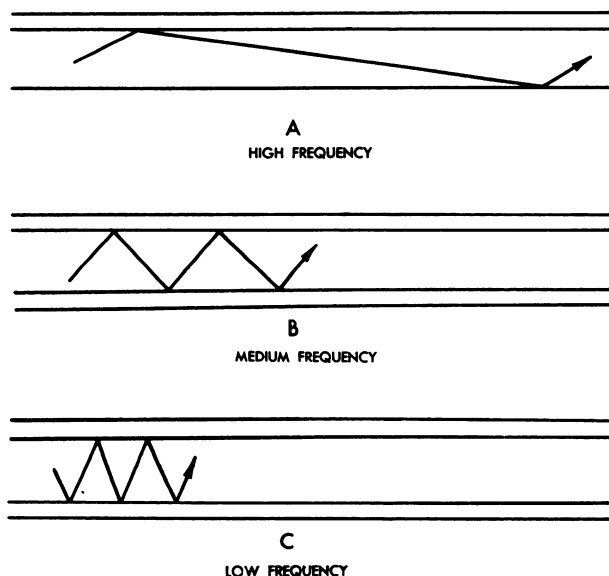


Figure 11-21. Angle at which Fields Cross Waveguide Varies with Frequency

zig-zag arrow at the velocity of light. But because of the long path, the wavefront actually travels very slowly along the waveguide. In Figure 11-21A the frequency is higher, and the wavefront or the group of waves actually travels a given distance in less time than those at C.

The axial velocity of a wavefront or a group of waves is called the *group velocity*. The relationship of the group velocity to diagonal velocity causes an unusual phenomenon. The velocity of propagation appears to be greater than the speed of light. As you can see in Figure 11-22, during a given time, a wavefront will move from point

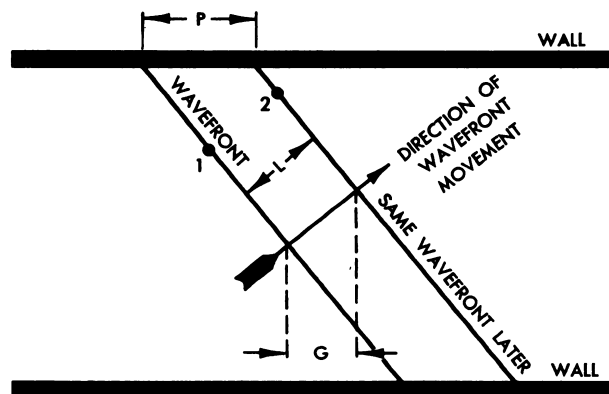


Figure 11-22. Relation of Phase, Group, and Wavefront Velocity

one to point two, or a distance  $L$  at the velocity of light ( $V_c$ ). Due to this diagonal movement (direction of the arrow), the wavefront has actually moved down the guide only the distance  $G$ , which is necessarily a lower velocity. This is called group velocity ( $V_g$ ).

But if an instrument were used to detect the two positions at the wall, they would be the distance  $P$  apart. This is greater than the distance  $L$  or  $G$ . The movement of the contact point between the wave and the wall is at a greater velocity. Since the phase of the r-f has changed over the distance  $P$ , this velocity is called the phase velocity ( $V_p$ ). The mathematical relationship between the three velocities is stated by the equation,

$$V_c = \sqrt{V_p V_g}$$

where

$V_c$  = Velocity of light =  $3 \times 10^8$  meters/second

$V_p$  = Phase Velocity

$V_g$  = Group Velocity.

This equation indicates that it is possible for the phase velocity to be greater than the velocity of light. As the frequency decreases, the angle of crossing is more of a right angle. In this condition the phase velocity increases. For measuring standing waves in a waveguide, it is the phase velocity which determines the distance between voltage maximum and minimum. For this reason, the wavelength measured in the guide will actually be greater than the wavelength in free space.

From a practical standpoint, the different velocities are related in the following manner: If the radio frequency being propagated is sine wave modulated, the modulation envelope will move forward through the waveguide at the group velocity, while the individual cycles of r-f energy will move forward through the modulation envelope at the phase velocity. If the modulation is a square wave, as in radar transmissions, again the square wave will travel at group velocity, while the r-f waveshape will move forward within the envelope.

Since the standing wave measuring equipment is affected by each r-f cycle, the wavelength will be governed by the rapid movement of the changes in r-f voltage. Since intelligence is conveyed by the modulation, the transfer of intelligence through the waveguide will be slower than



the speed of light, as is the case in other types of r-f lines.

Because of the way the fields are assumed to move across the waveguide, it is possible to establish a number of trigonometric relationships between certain factors. As shown in Figure 11-23, the angle that the wavefront makes with the wall (angle  $\theta$ ) is related to the wavelength and dimension of the guide and is,

$$\cos \theta = \frac{\lambda}{2B}$$

where  $\lambda$  is the wavelength in free space of the signal in the guide, and  $B$  is the inside wide dimension of the guide. The group velocity ( $V_g$ ) is related to the velocity of light ( $V_c$ ) as follows:

$$\frac{V_g}{V_c} = \sin \theta = \sqrt{1 - \left(\frac{\lambda}{2B}\right)^2}$$

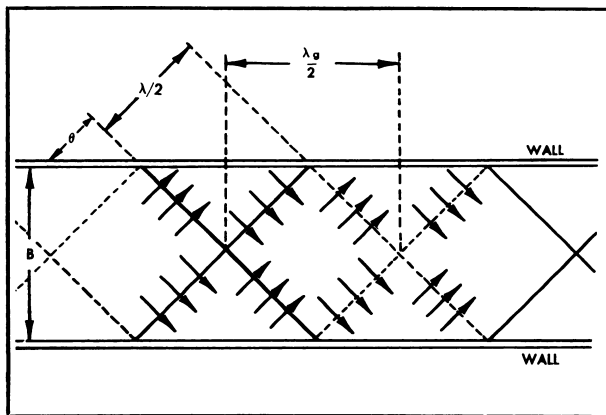


Figure 11-23. Trigonometric Relations Exist between Factors Indicated

Further, since it is possible to measure the wavelength in the guide ( $\lambda_g$ ), the wavelength in space is equal to,

$$\frac{\lambda_g}{\lambda} = \frac{1}{\sin \theta} = \frac{1}{\sqrt{1 - \left(\frac{\lambda}{2B}\right)^2}}$$

Solving this for  $\lambda$ , the equation becomes,

$$\lambda = \frac{2B\lambda_g}{\sqrt{\lambda_g^2 + 4B^2}}$$

After measuring the wavelength and the inside dimension of the waveguide, it is possible to calculate most other quantities associated with the waveguide (Figure 11-23).

## Numbering System of the Modes

The normal configuration of the electromagnetic field within a waveguide is called the *dominant* mode of operation. The mode developed for the rectangular waveguide, as was explained before, is the dominant mode of operation. The dominant mode for the circular waveguide was also shown in Figure 11-16. A wide variety of higher modes is possible in either type of waveguide. The higher modes in the rectangular waveguide are seldom used in radar, but some of the higher modes in the circular waveguide are useful.

For ease in identifying modes, any field configuration can be classified as either a transverse electric mode or a transverse magnetic mode. These modes are abbreviated TE or TM, respectively.

In a transverse electric mode, all parts of the electric field are perpendicular to the length of the guide and no E-line is parallel to the direction of propagation.

In a transverse magnetic mode the plane of the H-field is perpendicular to the length of the waveguide. No H-line is parallel to direction of propagation.

It is interesting to note from these definitions that the wavefront in free space or in a coaxial line is a TEM mode, since both fields are perpendicular to the direction of propagation. This mode cannot exist in a waveguide.

In addition to the letters TE or TM, subscript numbers are used to complete the description of

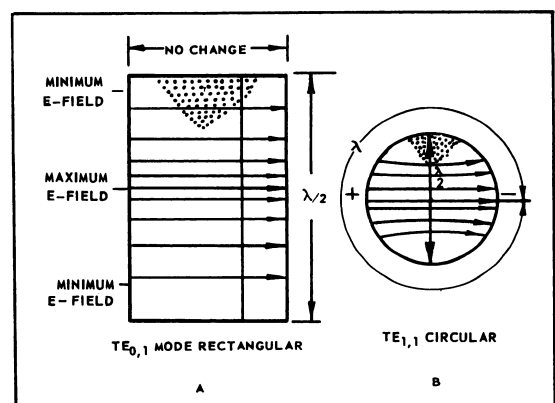


Figure 11-24. How to Count Wavelengths for Numbering Modes

the field pattern. In describing field configurations in rectangular guides, the first small number indicates the number of half-wave patterns that exist along the short dimension of the guide through the center of the cross section. The second small number indicates the number of transverse half-wave patterns that exist along the long dimension of the guide through the center of the cross section. For circular waveguides the first number indicates the number of full waves of the transverse field encountered around the circumference of the guide. The second number indicates the number of half-wave patterns that exist across the diameter.

### Counting Wavelengths for Measuring Modes

In the rectangular mode illustrated in Figure 11-24A, note that all the electric lines are perpendicular to the direction of movement. This makes it a TE mode. In the direction across the narrow dimension of the guide parallel to the E-line, the intensity change is zero. Across the guide along the wide dimension, the E-field varies from zero at the top through maximum at the center to zero on the bottom. Since this is one-half wave, the second subscript is one. Thus, the complete description of this mode is  $TE_{0,1}$ .

In the circular waveguide in Figure 11-24B, the E-field is transverse and the letters which describe it are TE. Moving around the circum-

ference starting at the top, the fields go from zero, through maximum positive (tail of arrows), through zero, through maximum negative (head of arrows), to zero. This is one full wave, so the number is one. Going through the diameter, the start is from zero at the top wall, through maximum in the center to zero at the bottom, one-half wave. The second subscript is one. The complete designation for the circular mode becomes  $TE_{1,1}$ .

Several circular and rectangular modes are possible. On each diagram illustrated in Figure 11-25 you can verify the numbering system. Note that the magnetic and electric fields are maximum in intensity in the same area. This indicates that the current and voltage are in phase. This is the condition which exists when there are no reflections to cause standing waves. In previous examples in which fields were developed, the fields were out of phase because of a short circuit at the end of the two-wire line.

### INTRODUCING FIELDS INTO A WAVEGUIDE

A waveguide, as was explained before, is a single conductor. Therefore, it does not have the two connections which ordinary r-f lines have, and it is necessary to use special devices to put energy into a waveguide at one end and to remove it from the other. In a waveguide, as with many other electrical networks, reciprocity

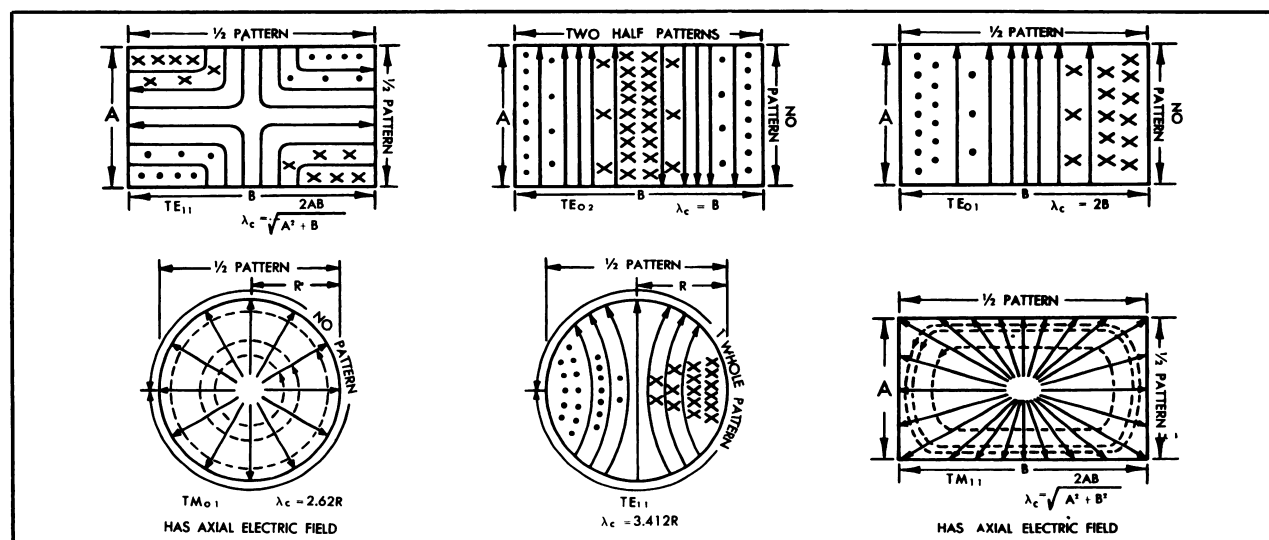


Figure 11-25. Various Modes in Waveguides

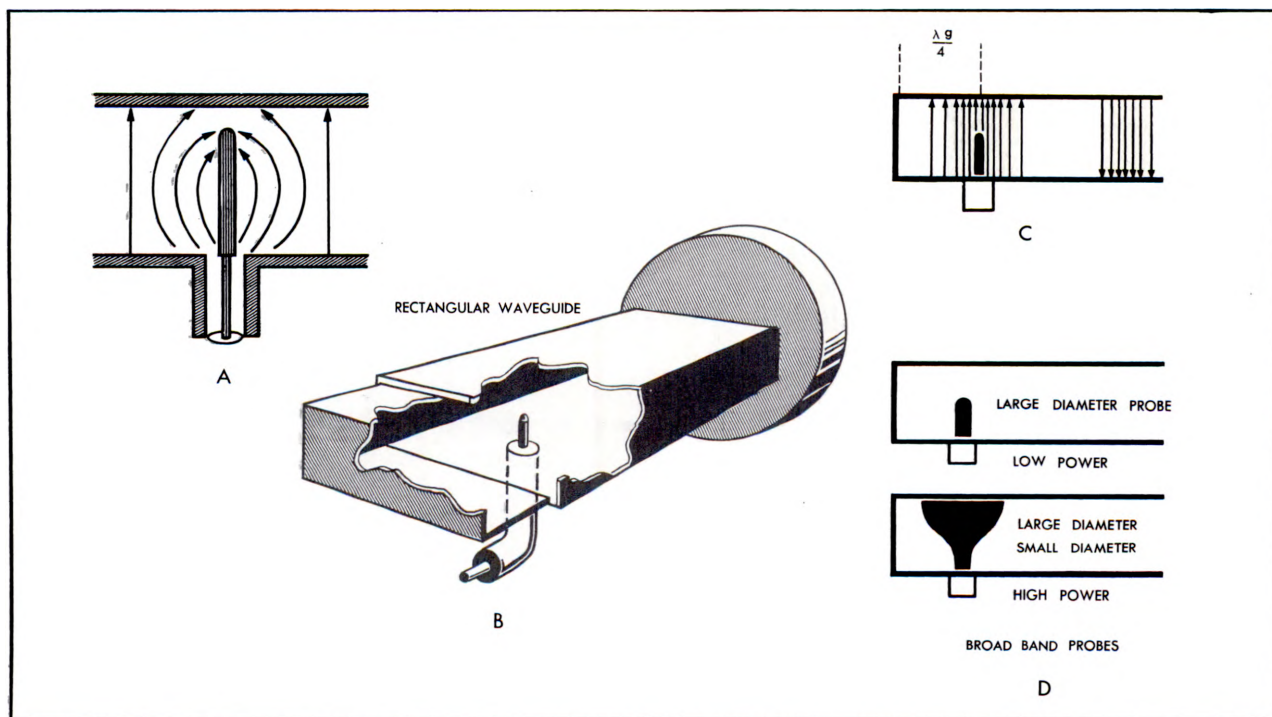


Figure 11-26. Exciting the Waveguide with Electric Field

exists in an excitation system — that is, energy may be transferred either to the waveguide or from the waveguide with the same efficiency.

Waveguides may be excited by three principal methods, namely, electric fields, magnetic fields, and electromagnetic fields.

#### Exciting with Electric Fields

When a small probe or antenna is placed in a waveguide and fed with an r-f signal, current will flow in the probe and set up an electrostatic field such as shown in Figure 11-26A. This causes the E-lines to detach themselves from the probe and to form in the waveguide. When the probe is located in the right place, a field having considerable intensity will be set up. The best place to locate the probe is in the center, parallel to the narrow dimension and one-quarter-wave-length away from the shorted end of the guide, as shown in Figure 11-26C.

Note here that the field is strongest at the quarter-wave point. This is the point of maximum coupling between the probe and the field. Of course the probe will work equally well in the center of any unidirectional field. For example, a three-quarter-wave distance from the

shorted end will also be a good spot to place the probe. Usually, the probe is fed with a coaxial cable. In comparison with the waveguide, this cable is extremely short, which insures that the greatest benefit will be derived from the waveguide.

The degree of excitation can be reduced by reducing the length of the probe, by moving it out of the center of the E-field, or by shielding it. Where it is necessary to vary the degree of excitation frequently, the probe is made retractable and the end of the waveguide fitted with a movable plunger. In airborne radar systems the position of the probe and the end piece is often predetermined by the factory and fixed permanently.

In pulse-modulated radar systems there are wide sidebands on each side of the carrier. In order that a probe feeding system does not discriminate too sharply against frequencies that differ from the carrier frequencies, *wide-band* probes are often used. This probe is large in diameter and is conical or door knob in shape. A conical probe is capable of handling high-powered signals.

The same kind of probe is used to take energy out of the guide and deliver it to the coaxial cable.

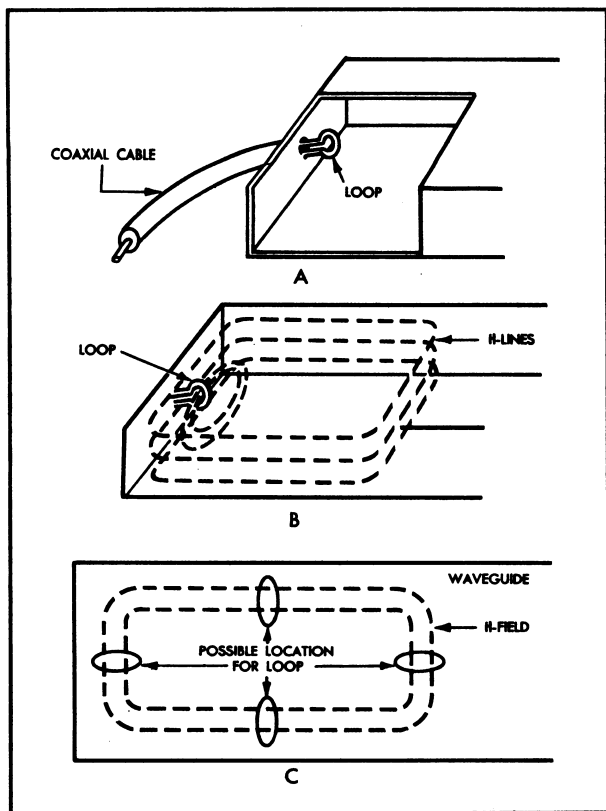


Figure 11-27. Excitation with Magnetic Fields

#### Excitation with a Magnetic Field

Another way of exciting a waveguide is by setting up a magnetic (H) field in the waveguide. This can be accomplished by placing a small loop which carries a high current in the waveguide. A magnetic field builds up around the loop. The field expands and fits the guide. If the frequency of the current is correct, energy will be transferred from the loop to the waveguide. A loop for transferring energy into a guide is shown at A and B in Figure 11-27. Notice that the loop is fed by a coaxial cable. The location of the loop for optimum coupling to the guide is at the place where the magnetic field that is to be set up will be of greatest strength. There are a series of places where this is true. Several are shown in Figure 11-27C.

When less coupling is desired, you can rotate or move the loop until it encircles a smaller number of lines of force.

When an excitation loop is used in radar equipment, its proper location is often predetermined and fixed during either construction or final tun-

ing at the factory. In test or laboratory equipment the loop is often made adjustable.

When a loop is introduced in a guide in which an H-field is present, a current will be induced in the loop itself. When this condition exists, the loop will take energy out of the waveguide as well as put energy into it.

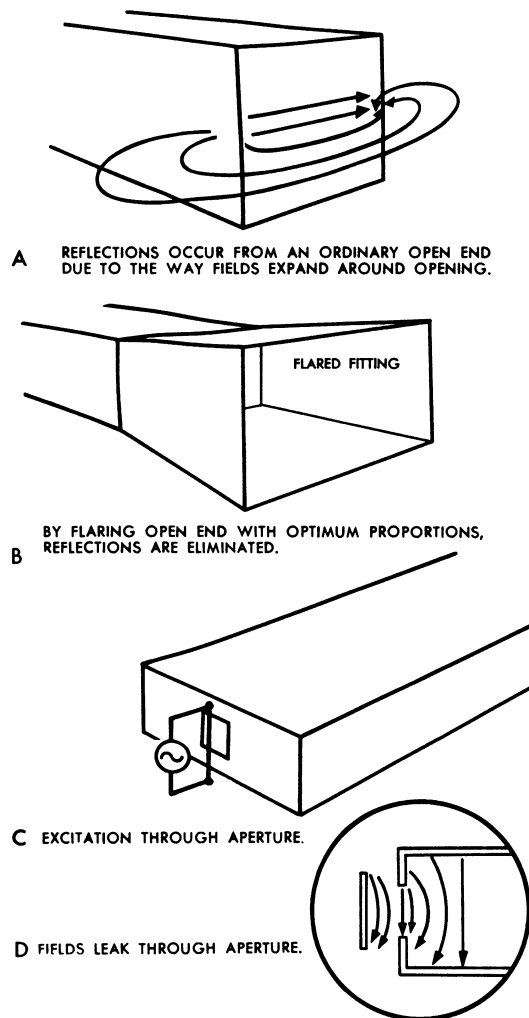


Figure 11-28. Excitation with Electromagnetic Fields

#### Excitation with Electromagnetic Fields

After learning what fields are like, you might think that a good way to either excite the waveguide or to let energy out of it is simply to leave the end open. But this is not the case, for when energy leaves a guide, fields form around the end of the guide and cause an impedance mismatch as shown in Figure 11-28A. In other



words, reflections and standing waves would result if the end were left open. Thus, simply leaving the end open is not an efficient way of letting energy out of the waveguide.

In order for energy to move smoothly in or out of a guide, the opening of the guide may be flared like a funnel as shown in Figure 11-28B. This makes the guide similar to a V-type antenna. The *funnel*, in effect, eliminates reflection by matching the impedance of free space to the impedance of the waveguide. When the mouth of the funnel is exposed to electromagnetic fields, they enter the funnel where they are gradually shaped to fit the waveguide. The funnel is directional in characteristic. It sends or receives the greatest amount of energy from in front of the opening.

Another method for either putting energy into or removing it from waveguides is through slots or openings. This method is sometimes used when very loose coupling is desired. In this method energy enters the guide through a small aperture, as you can see in Figure 11-28C. Any device which will generate an E-field may be placed near the aperture and the E-field will expand into the waveguide. A single wire is shown in Figure 11-28D. On it E-lines are set up parallel to the wire due to the voltage dif-

ference between different parts of the wire. The E-lines, in expanding, will exist first across the aperture, then across the interior of the waveguide. If the frequency is correct and the size of the aperture properly proportioned, energy will be transferred to the waveguide with a minimum of reflection.

## BENDS, TWISTS, JOINTS, AND TERMINATIONS

For energy to move from one end of a waveguide to the other without reflections, the size, shape, and dielectric material of the waveguide must be constant throughout its entire length. Any abrupt change in its size or shape results in reflections. Therefore, if no reflections are desired, any change in the direction or the size of the waveguide must be gradual. When it is necessary that the change in direction or size be abrupt, then special devices, such as bends, twists, joints, or terminations, must be used.

### Bends

Waveguides may be bent in several ways to avoid reflections (Figure 11-29). One is to make the bend gradual. It must have a radius of bend greater than two wavelengths in order to minimize any reflection. Some bends may be 90°.

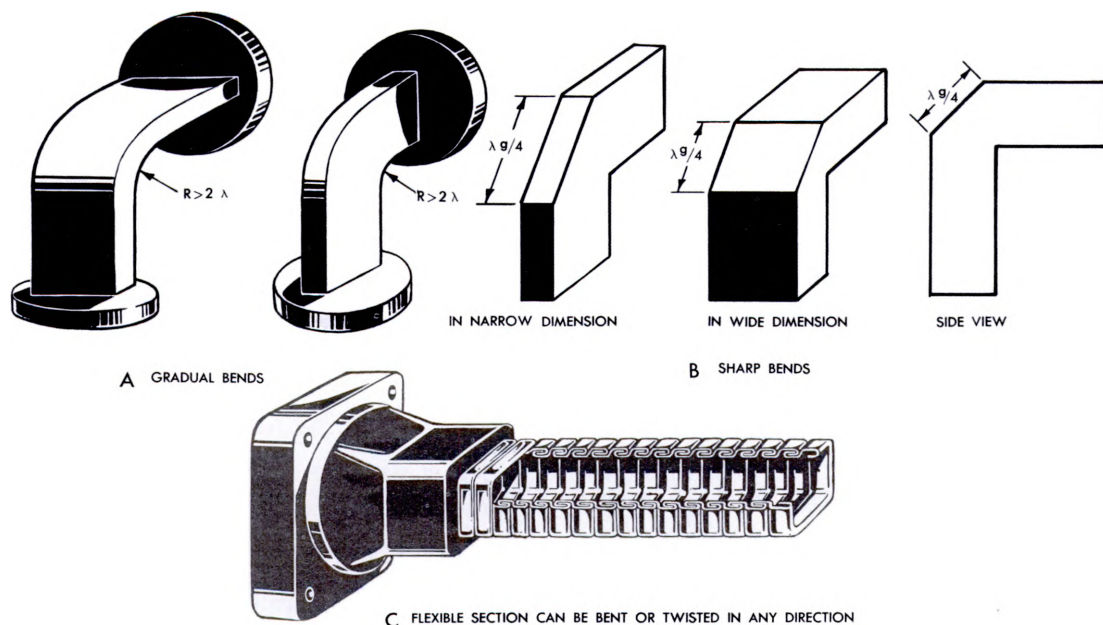


Figure 11-29. Types of Bends

Other bends may be greater or less than  $90^\circ$ , depending upon the requirements of the system. Still another type of bend is the sharp bend.

A bend can be made in either the narrow or the wide dimension of a guide without changing the mode of operation. In a sharp  $90^\circ$  bend, reflections will normally occur. To avoid this, the guide has two  $45^\circ$  bends, one quarter-wave apart. The direct reflection at one bend and the inverted reflection from the other bend will cancel and leave the fields as though no reflection had occurred.

To permit using any special bend which an installation might require, sections of a waveguide are often made flexible (Figure 11-29C). These sections can be bent or twisted in any desired direction. They consist of a spiral wound ribbon of brass. In cross section the winding is exactly the same size as a waveguide. The entire assembly is like a spiral spring in that it can be bent or twisted into any desired shape. As skin effect keeps the current at the inner surface of the waveguide, the inside surfaces of the flexible section are chromium-plated. This provides for maximum current conductivity. The outside of the section is covered with rubber. This gives the section flexibility and at the same time makes it both airtight and watertight.

### Rotating the Field

Sometimes it is desired to rotate the electromagnetic fields so that they are in the proper direction for matching. This may be accomplished by twisting the waveguide as shown in Figure 11-30. The twist should be gradual and extend over two wavelengths or more to prevent excessive reflections. Flexible sections also are used to rotate fields.

### Joints

Since it is impossible to mold an entire waveguide system in a radar set into one piece, it is necessary to construct it in sections and then

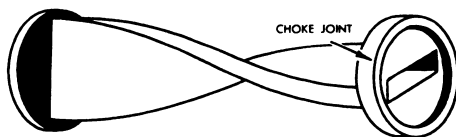


Figure 11-30. Twisted Section of Waveguide Rotates the Field with Minimum Reflection

to connect the sections together by joints. There are three main types of joints: these are the permanent, the semipermanent, and the rotating joints.

Joining two waveguide sections together requires more than that the sections be the same size and fit tightly at the joint. Irregularities at the joints set up standing waves and allow energy to escape. The permanent joint affords a good connection between the parts of a waveguide and has very little effect on the fields. This joint is made at the factory. When it is used, the waveguide sections are machined within a few thousandths of an inch and then welded together. The result is a hermetically-sealed and mirror-smooth joint.

Where it is necessary that sections be taken apart for normal maintenance and repair it is impractical to use a permanent joint. To permit portions of the waveguide to be taken apart, they are commonly connected together with semipermanent joints. The choke joint is the most common type of semipermanent joint. A cross-sectional view of a choke joint is shown in Figure 11-31A. It consists of two flanges which are connected to the waveguide at the center. The right-hand flange is flat, and the one at the left is slotted a quarter-wave deep at a distance a quarter-wave from the point where the walls of the guide are joined. The quarter-wave slot is shorted at the end. The two quarter-waves together become a half-wave and reflect a short circuit at the place where the walls are joined together.

Electrically, this creates a short circuit at the junction of the two waveguides. The two sections actually can be separated as much as a tenth of a wavelength without excessive loss of energy at the joint. This separation allows room to seal the interior of the waveguide with a rubber gasket for pressurization. The quarter-wave distance from the walls to the slot is modified slightly to compensate for the slight reactance introduced by the short space and the open circuit from the slot to the periphery of the flange.

The name choke joint is said to come from the similarity between the action of this joint on r-f fields and the action of an r-f choke in a power supply lead. An r-f choke keeps r-f in the circuit where it belongs. Similarly, the choke joint keeps the electromagnetic fields in the waveguide where they belong. The loss introduced by the

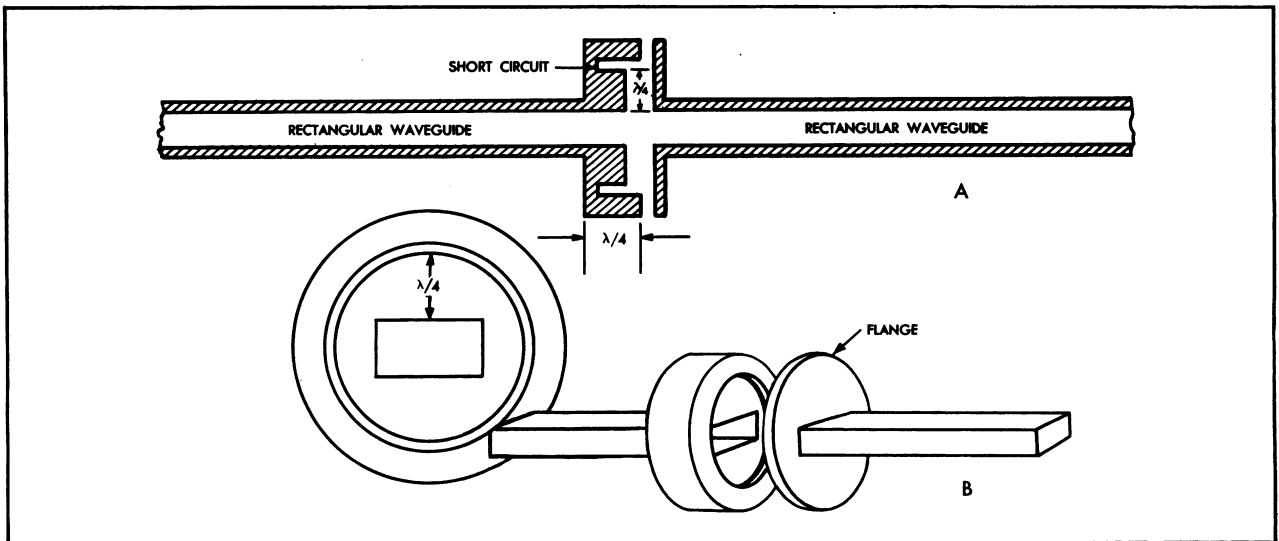


Figure 11-31. Choke Joint Keeps R-F Fields Inside Waveguide

well designed choke is less than .03 db, while an unsoldered permanent joint, well machined, has a loss of .05 db or more.

Rotating joints are usually required in a radar system where the transmitter is stationary and the antenna is rotatable. A simple method for rotating part of a waveguide system is by using a mode of operation that is symmetrical about the axis, as shown in Figure 11-32. This requirement is met by using a circular waveguide and a mode such as  $TM_{0,1}$ . In this method a choke joint may be used to separate the sections mechanically and to join them electrically. The waveguide rotates, but not the field. The con-

tinuous field produced by the fixed field minimizes reflections. This is basically the method used for all rotating joints. But since most radars use rectangular waveguide, the circular rotating joint must be inserted between two rectangular sections. A typical rotating joint is shown in Figure 11-33. The joint consists of two sections of circular guide, one fixed, the other rotating. At the ends of each section of circular guide is a transition from the circular to the rectangular waveguide.

Note that the rectangular guide is operating in the  $TE_{0,1}$  mode. (The lines in the illustration of the guide represent the E-lines.) The E-lines

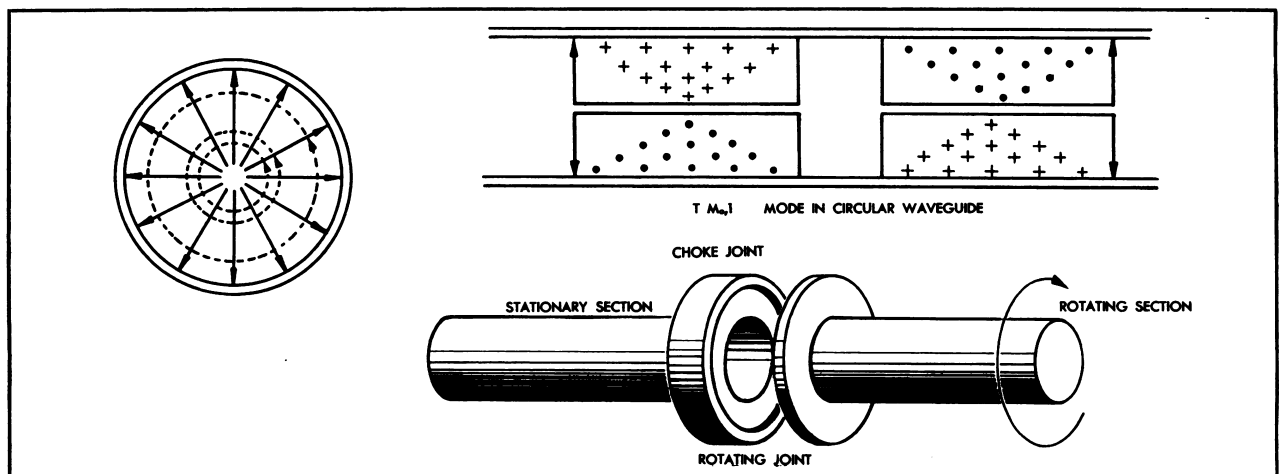


Figure 11-32. Rotating Joint and  $TM_{0,1}$  Mode in Circular Waveguide



couple from the rectangular guide into the circular guide and excite the circular guide in the  $TM_{0,1}$  mode. This is the mode that provides the required axial symmetry for rotating joints. At the top of the joint, the E-lines couple the energy into the rectangular guide that leads to the antenna, where the guide is operating in the  $TE_{0,1}$  mode.

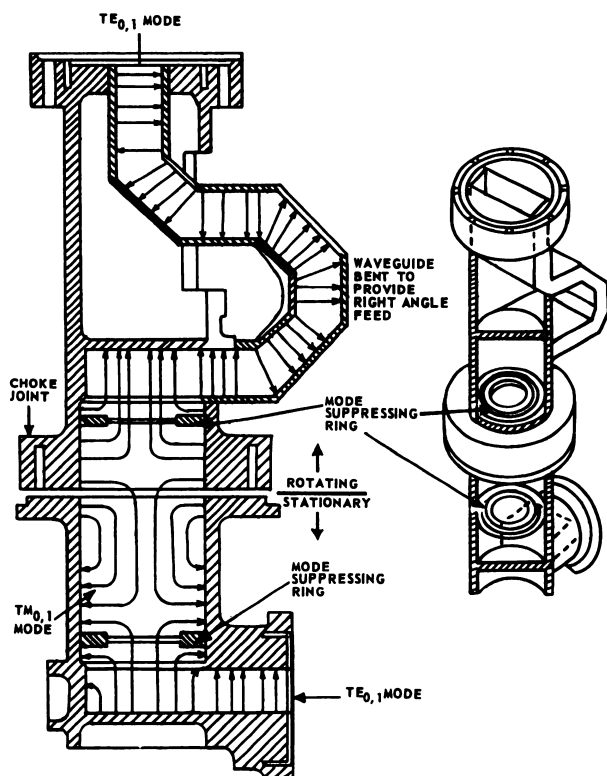


Figure 11-33. Rotating Joint with Rectangular Waveguide

Any circular guide that will carry the  $TM_{0,1}$  mode can also be excited in the  $TE_{1,1}$  mode. This mode is undesirable for a rotating joint. If you refer to Figure 11-25, you can see that the  $TE_{1,1}$  mode is not symmetrical with respect to the center of the guide. Thus, this mode would produce changes in the standing-wave ratio at the joint during rotation. For this reason, mode-suppressing rings are provided in the circular guide to suppress the  $TE_{1,1}$  mode. In addition, a teflon matching ring is used at each end of the joint to improve the impedance match between the circular and the rectangular waveguide sections.

## T-Junctions

Sometimes it is desirable to connect a section of waveguide into the side of another waveguide. This type of connection forms a T-junction. It may be connected either in the narrow side, as shown in Figure 11-34A, or in the wide side of the waveguide, as shown in Figure 11-34D. When the T-junction is in the plane of an H-field of a  $TE_{0,1}$  mode, it is called an *H-type* junction, and when the junction is in the plane of the E-lines, it is called an *E-type* junction.

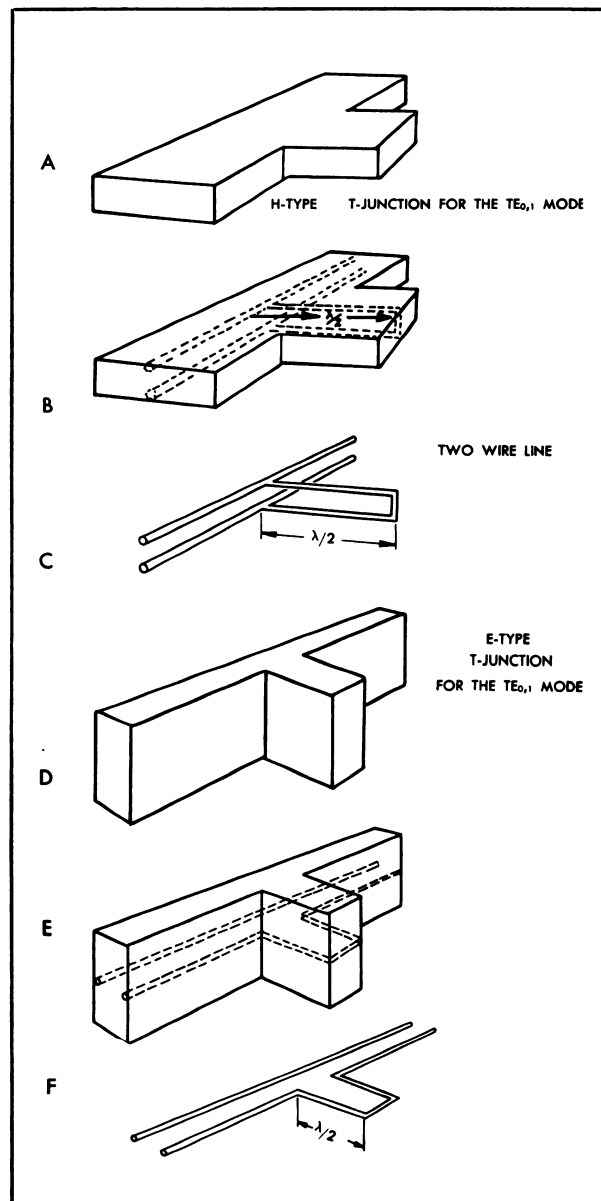


Figure 11-34. T-Junctions

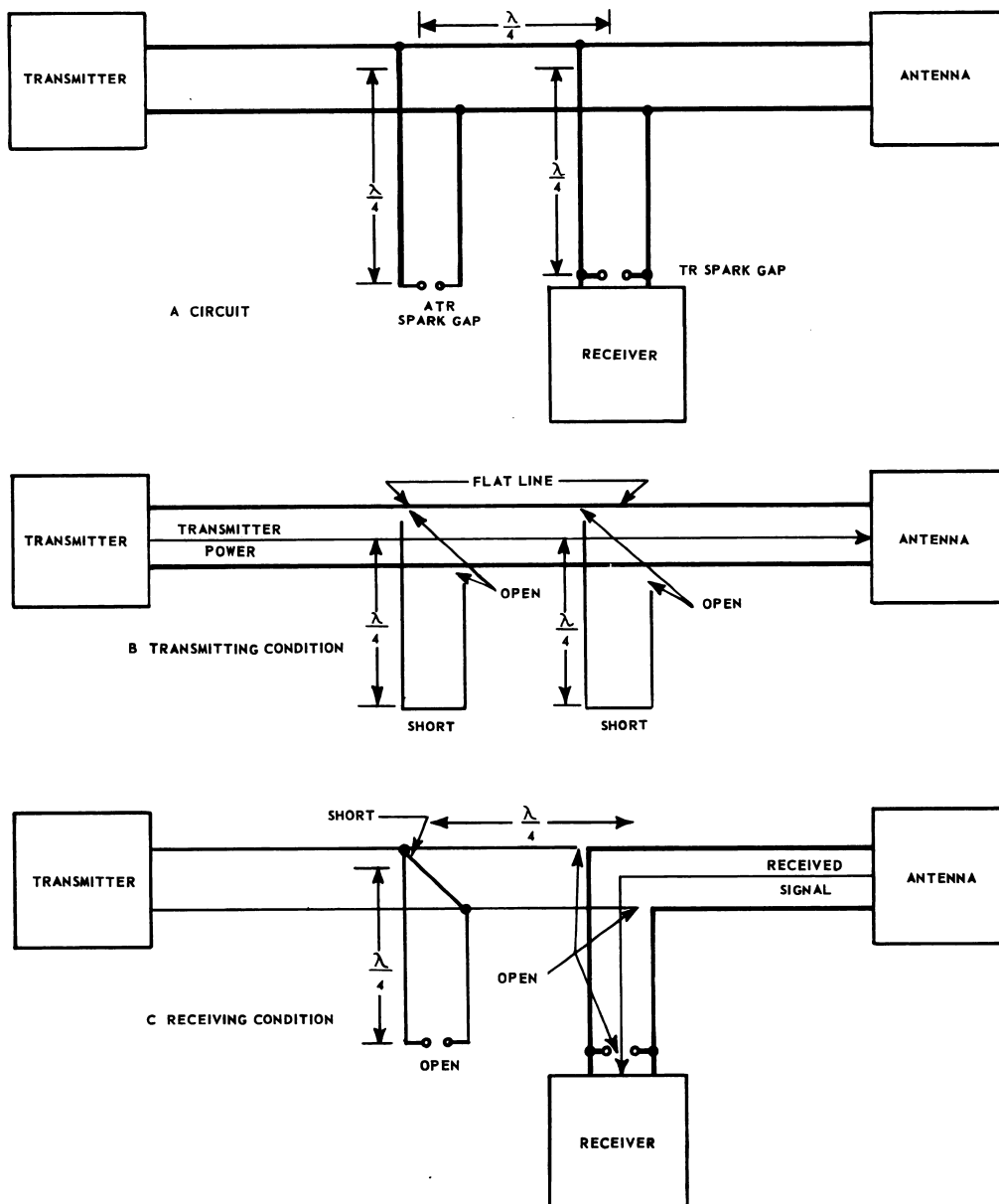


Figure 11-35. TR and ATR System

The H-type junction is effectively a parallel connection with the main line. For example, when the end of the T-joint shown in Figure 11-34B is short-circuited at a distance of a half-wave from the *center* of the waveguide, the result is the equivalent parallel circuit shown at C. This is a half-wave section that is connected to a wire line. This section will reflect a short circuit at the line and will not allow any energy to pass. Similarly, in the waveguide itself a short circuit is reflected to the center, where the E-

lines are supposed to be. Since an E-line cannot exist at a short circuit, no energy will pass that point. If the shorted end of the T-section were only a quarter-wave from the center of the waveguide, an open section would be reflected there and the passage of energy would be unaffected.

The E-type joint effectively is a series connection with one side of the main line, as you can see in Figure 11-34E. Its two-wire counterpart is shown at F. In this case, if the section added to the waveguide is a half-wave long, it

will act as a short at the junction, but will allow the energy to pass. The length of the added section will have to be a quarter-wave in order to open the circuit at the junction and to stop the ordinary flow of energy past it.

### **SIMULTANEOUS RECEPTION AND TRANSMISSION WITH COMMON ANTENNA**

Two separate antennas and transmission lines are sometimes used for transmitting and receiving r-f signals, one for each function. By using the same transmission line and antenna for both functions, the equipment and installation of equipment can be simplified to a large degree.

By modifying the transmission line, simultaneous reception and transmission using one antenna system can be accomplished. An important point to consider when the transmitter and receiver both operate on the same frequency, is keeping the transmitted signal out of the receiver. The high transmitter power could damage the sensitive receiver to a point where it could not receive at all.

#### **Parallel Duplexing Systems**

In radar where the receiver and transmitter use the same frequency, a duplexer system effects single antenna operation. The duplexer system contains TR and ATR circuits. The purpose of the TR circuit is to prevent the transmitter power from reaching the receiver. The purpose of the ATR circuit is to disconnect the transmitter during the period that the antenna is receiving.

Figure 11-35A shows a transmission line drawn as a two-wire line including the duplexer components just mentioned. The transmission line may be either a coaxial line or a waveguide; all operate on the same principle.

In Figure 11-35 the receiver transmission line is in parallel with the transmitter transmission line. A spark gap is placed in the receiver line a quarter-wavelength from the junction. This is the TR spark gap. One quarter-wavelength from the receiver line junction, toward the transmitter, an additional quarter-wavelength line is placed in parallel with the transmitter. This line is also terminated in a spark gap (ATR spark gap).

During transmission both spark gaps fire (Figure 11-35B) and cause the TR and ATR circuits to act as shorted quarter-wave transmission lines.

An open circuit, maximum impedance, is reflected at both junctions to the main transmission line. The transmitted power is conducted to the antenna without loss and none enters the receiver line.

During reception the amplitude of the received power is not sufficient to fire either gap. Under this condition the ATR circuit is now a quarter-wave transmission line terminated in an open. A short circuit is reflected back at the ATR junction with the transmission line, Figure 11-35C. This short circuit appears as an open circuit across the receiver line junction with the transmission line. The received signal, looking toward the transmitter, sees an open circuit and therefore enters the receiver circuit without loss.

Figure 11-36B illustrates the same circuit as Figure 11-35. The TR and ATR tubes are placed an odd number of quarter-wavelengths apart and a quarter-wavelength from the waveguide. When the transmitter fires, both tubes fire. At A and B in the waveguide sections, Figure 11-36B, an open circuit is reflected. This permits the transmitted power to be conducted to the antenna without loss. Since a short is across the TR tube, the transmitted power does not enter the receiver.

During reception neither the TR nor the ATR tube fires. The ATR section of waveguide is now an open quarter-wave stub, which has the same effect as a shorted half-wave section. Therefore, it reflects a short to point A. Since the distance between points A and B is a quarter-wave, the short at point A is reflected as a high impedance at point B. This effectively breaks the line from the receiver junction toward the transmitter, and prevents received energy from entering the transmitter and being dissipated. Therefore, the received signal passes into the receiver.

#### **Series-Parallel Duplexer**

The action of the series-parallel duplexer is similar to the parallel duplexer system with the exception that the ATR is placed in series with the transmitter while the TR is placed in parallel with the transmitter. Figure 11-37 illustrates the series-parallel duplexer.

The ATR circuit, explained under ATR Devices, is in series with the transmitter, and is placed one wavelength from the junction of the

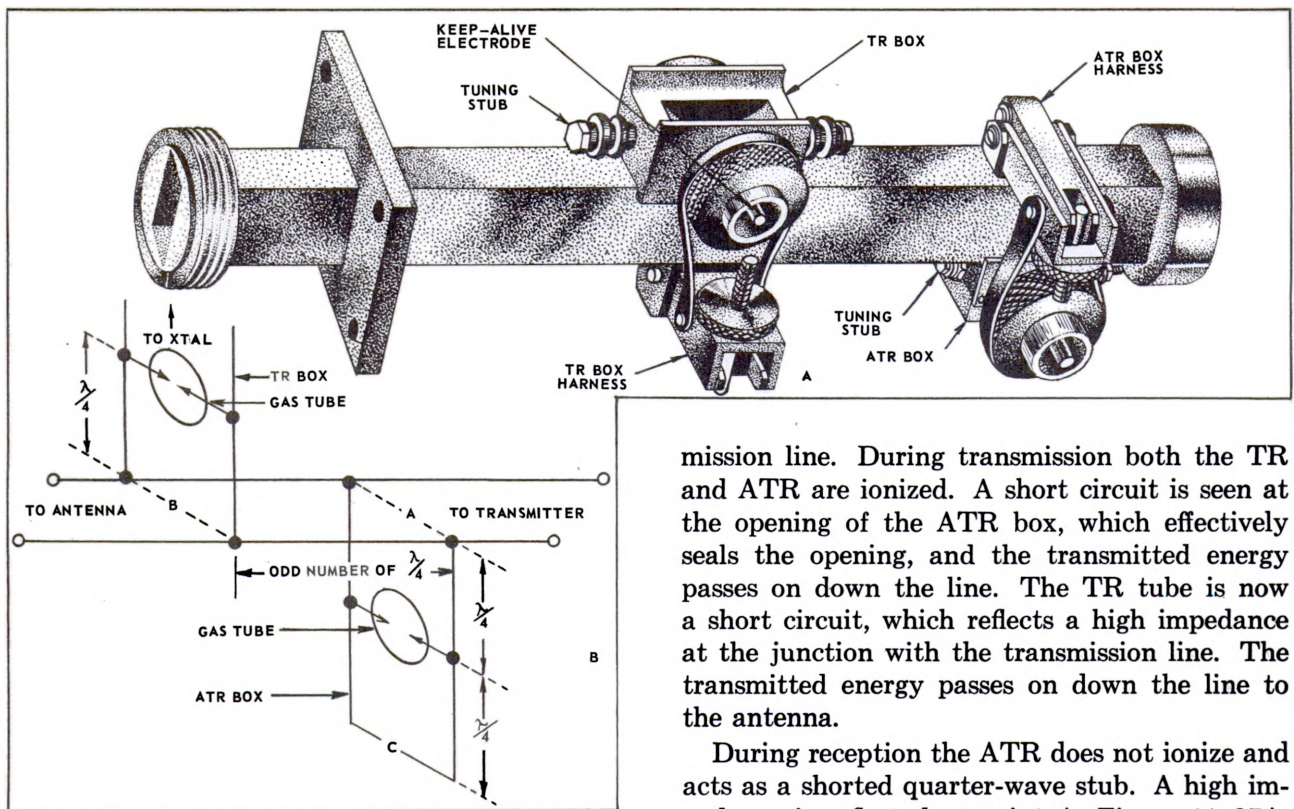


Figure 11-36. TR and ATR System in Waveguide

mission line. During transmission both the TR and ATR are ionized. A short circuit is seen at the opening of the ATR box, which effectively seals the opening, and the transmitted energy passes on down the line. The TR tube is now a short circuit, which reflects a high impedance at the junction with the transmission line. The transmitted energy passes on down the line to the antenna.

During reception the ATR does not ionize and acts as a shorted quarter-wave stub. A high impedance is reflected at point A, Figure 11-37A. As the distance from point A to point B is a full wave, a high impedance is reflected at point B. The received energy is prevented from entering the transmitter section and therefore passes into the receiver section.

TR circuit. The TR tube, as before, is a quarter-wavelength from the junction with the trans-

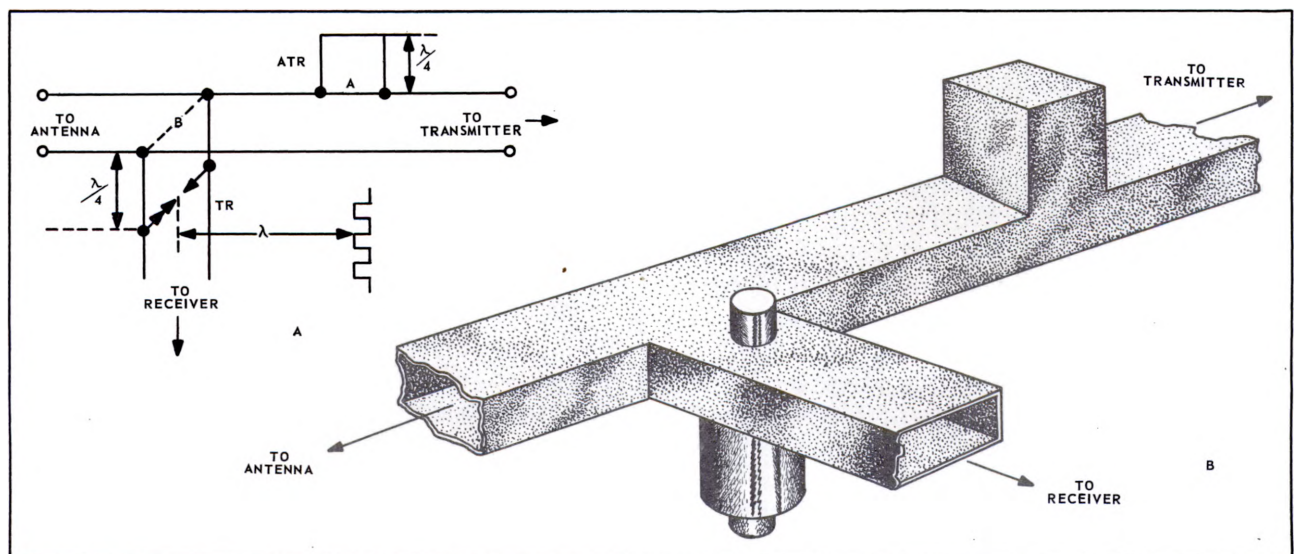


Figure 11-37. Series-Parallel Duplexer



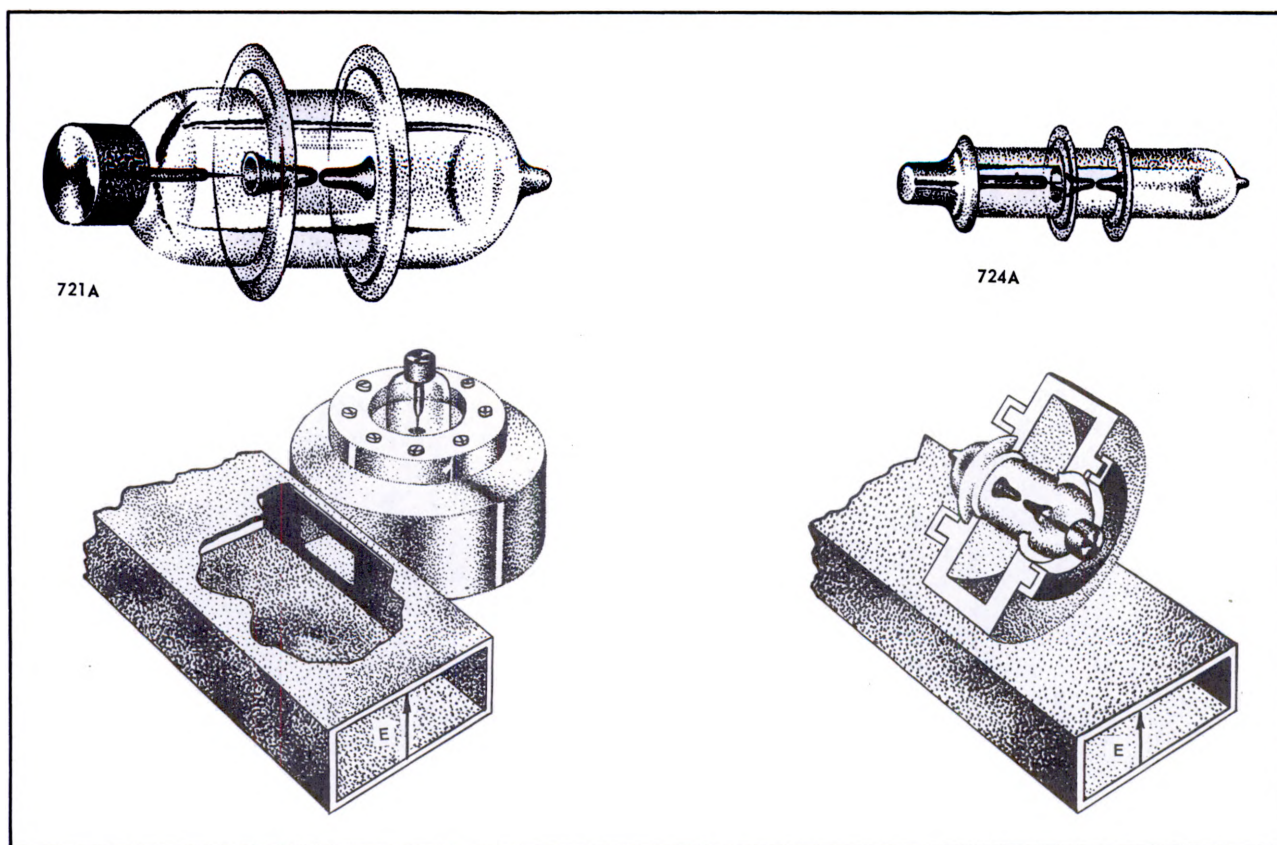


Figure 11-38. TR Tubes

### TR Tubes

A spark gap is not ideal in its operation. It acts as a high impedance when not fired; when fired, it acts as a nonlinear resistance with a low voltage across itself. The characteristics of a spark gap change with use. To increase its efficiency, the spark gap is enclosed in a partially evacuated glass envelope and is connected to the transmission line through a resonant cavity. Figure 11-38 illustrates the TR tube and the connection to the waveguide.

The purpose of the TR tube is to prevent the transmitted energy from entering the receiver. If the transmitted energy entered the receiver, the crystal mixer of the receiver would be damaged severely or would be completely burned out. To protect the crystal mixer, the TR tube must fire, or ionize, immediately after the transmitter fires.

The majority of TR tubes are filled with a mixture of argon gas and water vapor. To aid them in ionization, a negative voltage is applied

to an electrode that is placed near one of the gap electrodes. A TR tube showing the keep-alive electrode is illustrated in Figure 11-39.

The negative voltage maintains a steady glow discharge that provides a continuous supply of ions and free electrons. When r-f energy from the transmitter is applied to the electrodes, the

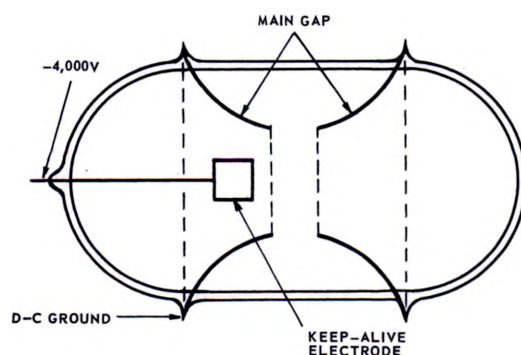


Figure 11-39. TR Tube Showing Keep-Alive Electrode



tubes fire more quickly. In this manner the keep-alive electrode acts as a cathode, while the back of one of the other electrodes acts as the anode. There are different requirements for different radars, but a typical example is that the TR must fire or ionize within .01 microsecond after application of the transmitted pulse. By the same token the tube must deionize 6 microseconds after termination of the transmitted pulse. This interval is called *recovery time*.

To prevent the crystals in the receiver from being damaged when the radar set is turned off, some radar sets are provided with a crystal shutter or gate. This normally consists of a metal vane across the waveguide, which is automatically positioned when the set is turned off.

### ATR Devices

When an ATR tube is used in parallel with the transmitter, it is generally of the same construction as the TR tube, but it has no keep-alive voltage applied. When the ATR device is used in series with the transmitter, it is of a different construction.

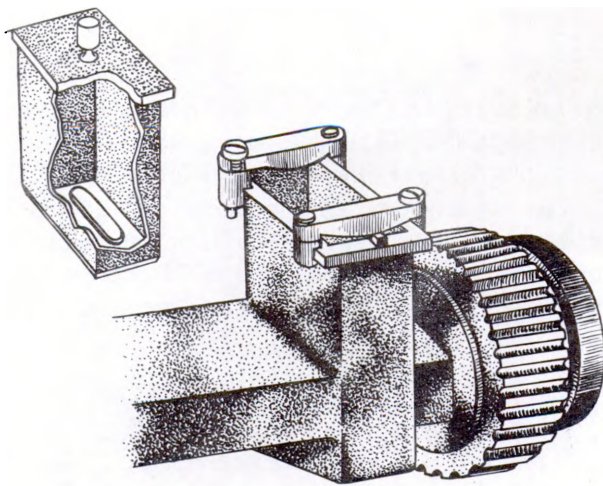


Figure 11-40. IB35 ATR Box and Mounting

Figure 11-40 illustrates the IB35 ATR box, which actually replaces a portion of the broad side of the waveguide. The ATR box consists of a cavity that is filled with argon gas and water vapor. During transmission the gas within the box ionizes and in effect completes the section of the waveguide that it replaces.

### Polarization-Shifting Duplexer

One type of duplexer used on a number of radar equipments is termed a polarization-shifting duplexer (see Figure 11-41). In this particular type of duplexer the rectangular waveguide at the antenna end of the duplexer is oriented with its transverse dimensions  $90^\circ$  from those at the end of the duplexer nearest the transmitter. The receiver arm is oriented with its electric vector parallel to the electric vector in the waveguide through which the received signal arrives at the antenna.

Both transmitted and received signals are carried in rectangular waveguide and arrive at the duplexer with a  $TE_{0,1}$  mode of transmission, that is, with the electric vector in the rectangular waveguide at all times parallel to the short dimension of the waveguide cross section. Energy being propagated toward the antenna on transmission has its polarization shifted  $90^\circ$  in passing through the duplexer and readily continues in the  $TE_{0,1}$  mode through the waveguide to the antenna. It does not couple into the receiver arm, except to a limited degree, because the polarization of electric vectors in the transmitted wave and the receiver arm differ by  $90^\circ$ . The energy of the received wave is coupled into the receiver arm but cannot pass through to the transmitter because of the transverse orientation of the waveguide at that end and because of the ATR device.

The duplexer consists of a section of circular waveguide which is converted at each end to the rectangular section waveguide by transition sections. Within the duplexer (that is, the cylindrical section) 20 gas-filled quartz tubes are placed at regular intervals along the axis of the cylinder. Each is positioned on a diameter of the cylinder. Four of these tubes (the first four encountered from the transmitted signal input end) are all parallel to the long dimension of the rectangular section waveguide at that end. These are vertical in Figure 11-41. The fifth quartz tube is placed at an angle of  $5.63^\circ$  counterclockwise from the first four (viewed from the transmitter end). Each of the remaining 15 tubes is turned an additional  $5.63^\circ$  CCW, so that the last tube is at right angles to the first four.

The transmitted wave enters the duplexer transition in the  $TE_{0,1}$  mode; the electric field is parallel to the short dimension of the waveguide. Since the duplexer is circular in cross section, the



wave proceeds in the  $TE_{1,1}$  mode, but the orientation of the transmitted energy is maintained, assisted by transverse shorting bars inside the transition section, near the end of the duplexer. Just beyond the branch arm for the received signal the transmitted energy strikes the first of the gas-filled quartz tubes. These tubes are filled with argon, generally at a pressure of about 20 mm. The high-power r-f energy ionizes the argon molecules, breaking down the initial resistance of the tube and causing it to conduct. This forms a very low impedance path across the cylindrical duplexer, and the first four "in-line" tubes do effectively the same thing as the

metal shorting rods. They tend to maintain the existing polarization of the transmitter wave.

The high-power wave now meets the fifth tube, which has an angular displacement of  $5.63^\circ$  from the first four. The direction of the E vectors is shifted by this amount, and the rotation will continue through the 16 tubes until the transmitted wave arrives at the twentieth tube. Polarization has been shifted through a full  $90^\circ$ , or is at right angles to that of the energy entering the duplexer. The net steady-state result of field interaction within the "fired" duplexer is a nearly uniform axial conducting surface twisting through  $90^\circ$ . The purpose of the matching slug

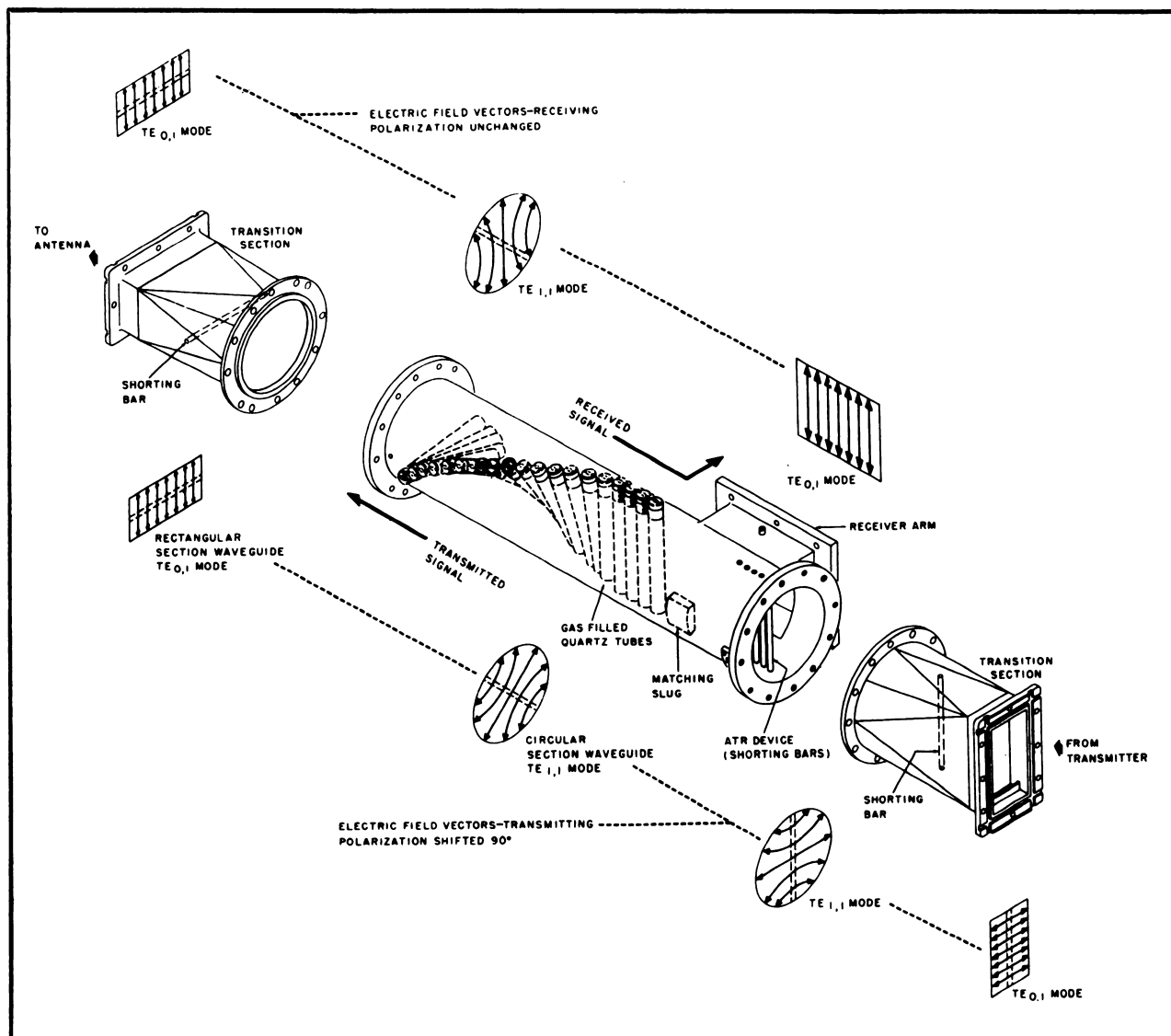


Figure 11-41. Polarization-Shifting Duplexer

on the side of the duplexer opposite the opening into the receiver arm is to compensate for the effect upon the transmitted wave of the discontinuity in the walls of the duplexer caused by the opening into the receiver arm.

The received signal from the antenna arrives at the duplexer with a polarization determined by the orientation of the rectangular waveguide through which it has passed. It is too weak to fire the gas-filled quartz tubes placed spirally across the duplexer. They do not function as in transmission and the received wave passes through with its polarization unchanged.

For two reasons it is coupled into the receiver arm. First, the orientation of the rectangular waveguide of the receiver arm is aligned with the polarization of the incoming wave, and the latter can therefore proceed into the receiver arm without serious attenuation. Second, the received wave is prevented from passing directly through the duplexer toward the transmitter by the five shorting bars across the duplexer just beyond the receiver arm. These serve to reflect the received signal and to set up standing waves with a current minimum at the receiver branch leading to the TR cavity. This provides maximum coupling of the received signal into the TR cavity. These shorting rods are, therefore, an ATR device. Figure 11-41 is of a typical polarization-shifting duplexer.

## Matching Devices

Some devices which are used in radar introduce inductance or capacitance. Sometimes these reactances are deliberately introduced. When they are present and not desired, they can be tuned out with small fins or plates in a waveguide.

When you couple energy into or out of a waveguide, you should effect a proper impedance match between the generator, the coupling elements, the waveguide, and the load. In this respect, the waveguide acts exactly like a two-wire line and follows the same principles outlined previously. In other words, unless you terminate the waveguide in its characteristic impedance, standing waves are created.

You have used appropriate matching elements such as impedance-matching transformers and reactive elements to effect maximum transfer of energy. While the same principles are used in matching that were described earlier, the physical appearance of the matching elements is somewhat different in waveguides.

Figure 11-42 shows a number of reactive plates that are used deliberately to introduce capacitance or inductance in a waveguide. When these plates are used as shown in Figure 11-42A, they set up oscillations in the higher modes. Since a waveguide is too small for higher modes at the same frequency, these frequencies are not

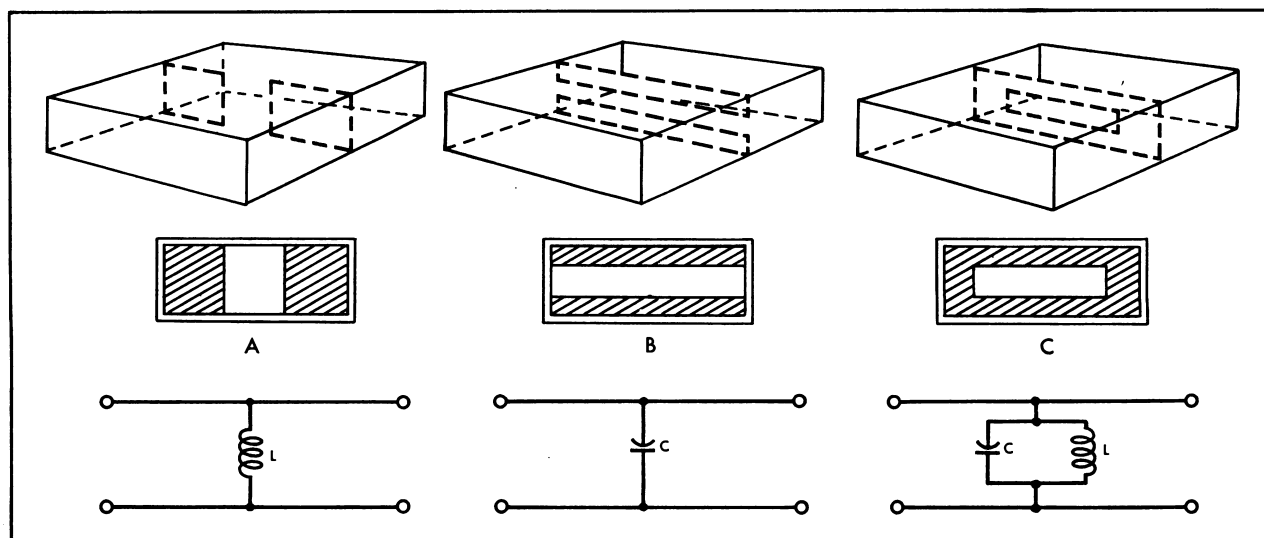


Figure 11-42. Reactive Plates in Waveguide

propagated, but remain in the vicinity of the plates. If the edges of the plates are vertical with respect to the plane of the H-field, the modes produced are the TM type. The effect of this on power flow is that of inductance across the two-wire line. This causes reflections and a shift in the standing wave pattern. The wider the space between the plates, the greater the inductive reactance.

When the partitions are arranged perpendicularly to the E-field as in Figure 11-42B, a local E-field and the higher modes of oscillation are set up between the edges of the plates. These oscillations cannot be propagated but do change the dominant mode to a TE mode and introduce capacitive reactance. As with the TM mode, the wider the opening, the greater the reactance.

From these facts it would seem that combining both types of plates and leaving a small opening in a large guide as in Figure 11-42C would produce a resonant circuit. This is approximately true, if the dimensions are correct. At resonance a resonant circuit acts as a high resistance. In this condition a small opening would introduce a high shunt resistance, and the guide would in effect have connected across it a resonant circuit.

**POST OR TUNING SCREW.** Inserting a metallic element which is either cylindrical in shape or in the form of a strip will produce a shunt reactance across a waveguide. When the element extends fully across the narrow part of the waveguide, it is called an inductive post (Figure 11-43A). When the post is in the form of a screw so that the insertion depth can be varied, as in Figure 11-43B, it is called a tuning screw.

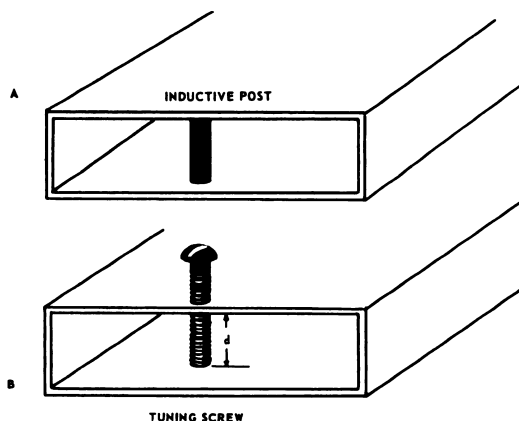


Figure 11-43. Waveguide Reactances

The reactance obtained by the tuning screw varies with the depth of insertion  $d$ . When  $d$  is small, the screw will act as a shunting capacitance. When  $d$  is increased, a point will be reached where  $d$  becomes equal to a quarter-wavelength. Resonance occurs at this point and the reactance is zero. If  $d$  is increased beyond this point, the screw becomes inductive. Since the short length of the guide is usually greater than a quarter-wavelength, the element is inductive when it extends fully across the waveguide. In this instance the element is referred to as an inductive post (Figure 11-43A).

**STUB REACTANCE.** Open or short-circuited stubs are also used to obtain reactances in waveguides. Shorted stubs are generally used to prevent undesired radiation of energy. By placing the stub as shown in Figure 11-44A, it acts as

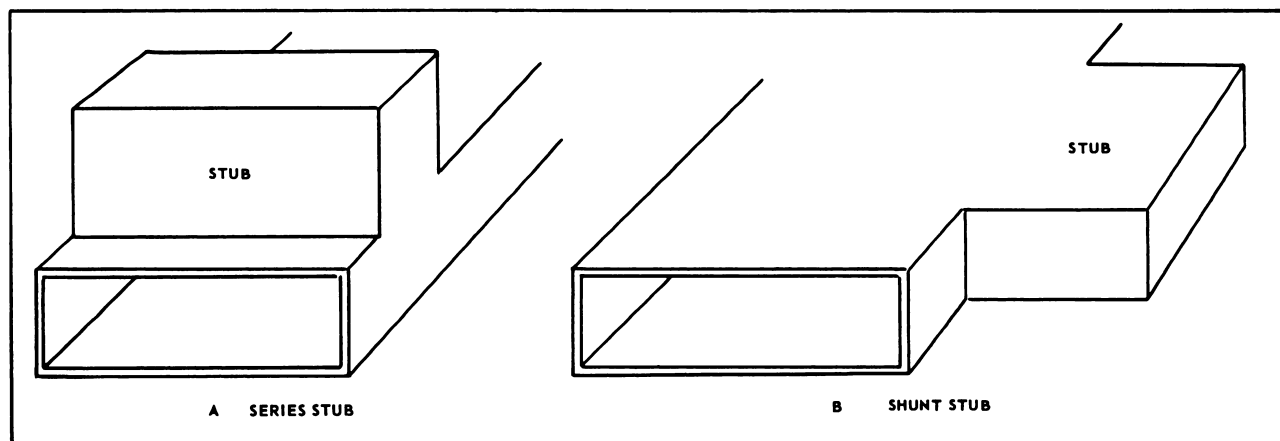


Figure 11-44. Waveguide Stubs

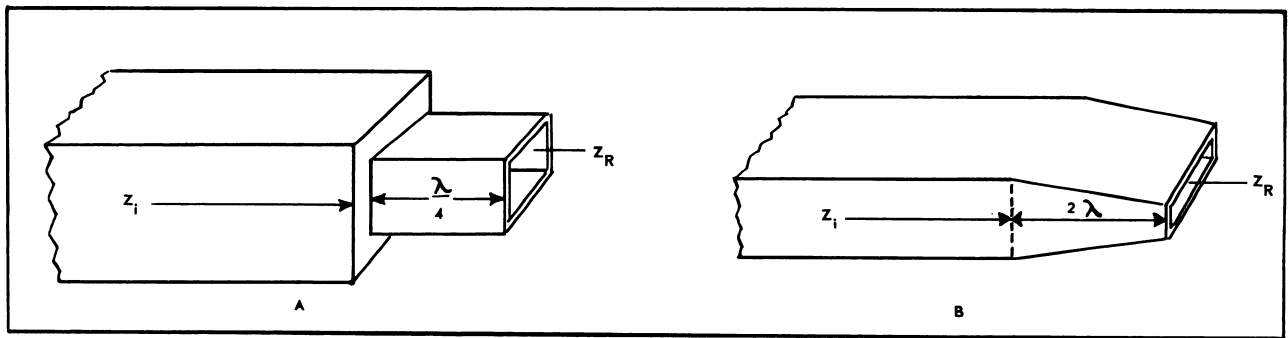


Figure 11-45. Waveguide Transformers

an impedance in series with the line and is therefore called a series stub. Figure 11-44B shows a stub acting as a shunt impedance across the waveguide. For this reason it is called a shunt stub.

**IMPEDANCE TRANSFORMERS.** You know that an impedance transformer is used to change one value of resistance to another in order to obtain maximum transfer of power. In a waveguide the transformer is placed at a point where the impedance is pure resistance but not equal to  $Z_0$ . Placing the transformer at this point transforms this value of resistance to  $Z_0$  and thus matches the waveguide to the load. A *quarter-wave transformer* can be constructed by using a section of waveguide whose dimensions are either greater or smaller than those of the main waveguide. This is illustrated in Figure 11-45A.

Figure 11-45B illustrates another type of transformer, the *tapered-line* transformer. In this type of transformer the dimensions of the guide are varied very gradually, thus changing its characteristic impedance from that of the main line to that of the load. When this change in impedance is made very gradually, the impedance transformation is effective over a very wide band of frequencies. Note that the tapered section is longer than two wavelengths. In comparison, the quarter-wave transformer is effective only at certain frequencies, those at which the quarter-wave transformer is of the proper length.

#### Terminating a Waveguide

Since a waveguide is a single conductor, it is not as easy to define its characteristic impedance ( $Z_0$ ) as it is for a coaxial line. Nevertheless, you can think of the characteristic impedance of a

waveguide as being approximately equal to the ratio of the strength of the electric field to the strength of the magnetic field for energy traveling in one direction. This ratio is equivalent to the voltage-to-current ratio in coaxial lines on which there are no standing waves.

The lowest characteristic impedance of a circular waveguide is about 350 ohms. In a rectangular waveguide,  $Z_0$  may be any value, depending on the dimensions of the waveguide and the frequency of the electrical energy. In the waveguide,  $Z_0$  is directly proportional to the narrow dimension when the other dimension and the frequency are fixed and may vary from approximately zero to 465 ohms.

On a waveguide there is no place to connect a fixed resistor to terminate it in its characteristic impedance as there is on a coaxial cable. But there are a number of special arrangements that accomplish the same thing. One consists of filling the end of the waveguide with graphited-sand as shown in Figure 11-46A. As the fields enter the sand, currents flow in it. These currents create heat, which is instrumental in dissipating energy. None of the energy thus dissipated as heat is reflected back into the guide. Another arrangement, Figure 11-46B, uses a high resistance rod, which is placed at the center of the E-field. The E-field (voltage) causes current to flow through the rod. The high resistance of the rod dissipates the energy as an  $I^2R$  loss.

Still another method for terminating a waveguide is to use a wedge of high resistance material, Figure 11-46C. The plane of the wedge is placed perpendicular to the magnetic lines of force. When the H-lines cut the wedge, a voltage is induced in it. The current produced by the

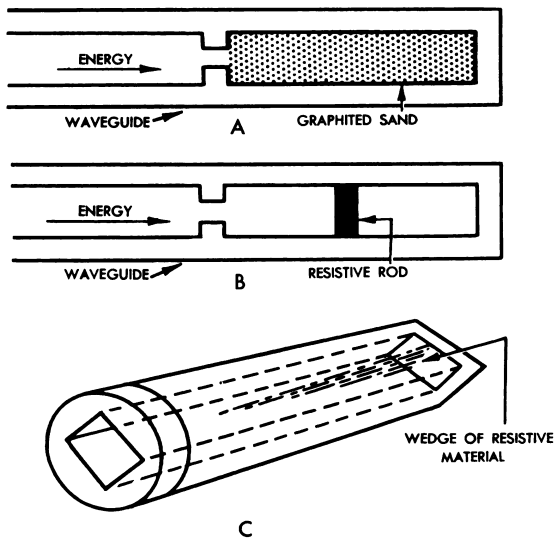


Figure 11-46. Termination for Minimum Reflections

induced voltage flowing through the high resistance of the wedge creates an  $I^2R$  loss. This loss is dissipated in the form of heat. This permits very little energy to reach the closed end to be reflected.

Each of the preceding terminations is designed to match the impedance of the guide in order to insure a minimum of reflection. On the other hand, there are many instances where it is desirable for all the energy to be reflected from the end of the waveguide. The best way to accomplish this is to permanently weld a metal plate at the end of the waveguide as shown in Figure 11-47A.

When it is necessary that the end be removable, a removable end plate is attached to the end of the guide. For this method to be satisfactory, the contact between the guide and the end plate must be exceptionally good in order that the H-field will not be attenuated when current flows. Perfect contact is not required when the connection is made at a point of minimum current. This point is located at a quarter-wave from the end. When connection is made at this point, a cup is used instead of the end plate, Figure 11-47B. This cup is a quarter-wave long and large enough to fit over the end of the guide. The voltage between opposite sides of the cup opening is high, but as the reflected H-field cancels the incident H-field, the resulting current is very small and reflection is at a minimum.

When the end must be adjustable, the contact must be nearly perfect. However, it is impossible to get a perfect contact. The best arrangement, one which is similar to the choke joint previously explained, consists principally of an adjustable plunger that fits into the guide as shown in Figure 11-47C. The walls of the waveguide and the plunger form a half-wave channel. The half-wave channel is closed at the end and reflects a short circuit across the other end, where a perfect connection is supposed to exist between the wall and the plunger. The actual physical contact is made at a quarter-wave distance from the short circuit, where the current is minimum due to the standing waves. This makes it possible for the plunger to slide loosely in the guide at a point where the contact resistance to current flow is very low.

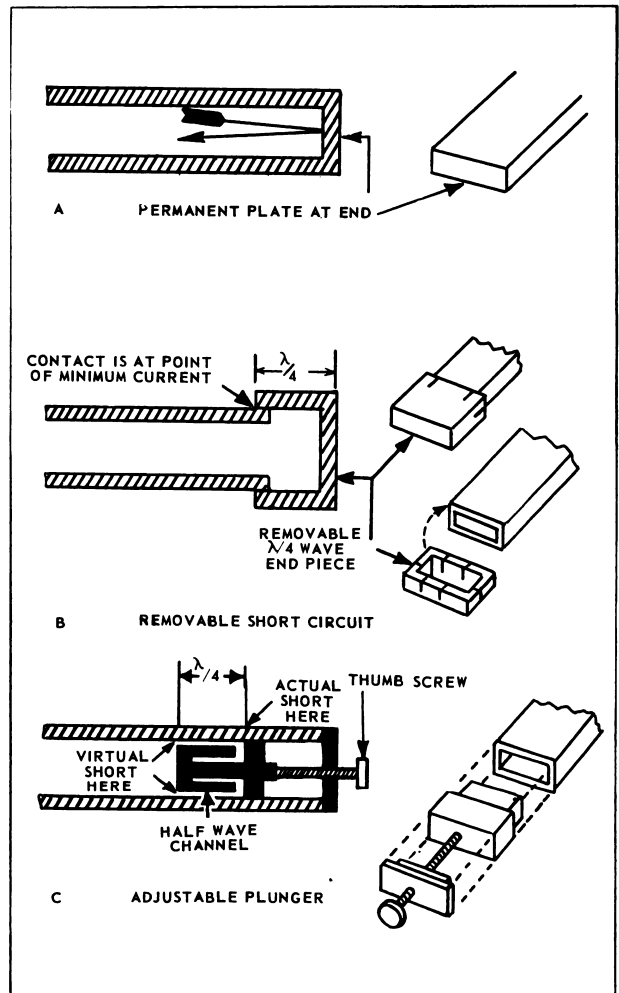


Figure 11-47. Termination for Maximum Reflection

## CAVITY RESONATORS

In ordinary radio work the conventional low-frequency resonant circuit consists of a coil and capacitor that are connected either in series or in parallel as shown in Figure 11-48A. To increase the resonant frequency, it is necessary to decrease either the capacitance or the inductance, or both. However, a frequency is reached where the inductance is a half-turn coil and where the capacitance consists only of the stray capacitance in the coil. In this circuit the current handling capacity and breakdown voltage for the spacing would be low.

The current carrying ability of a resonant circuit may be increased by adding half-turn loops in parallel. This does not change the resonant frequency appreciably because it adds capacitance in parallel — which lowers the frequency, and inductance in parallel — which increases the frequency. As the effects of one cancel the effects of the other, the frequency remains about the same.

In Figure 11-48C several half-turn loops are added in parallel. D shows several parallel quarter-wave Lecher lines, which you recall are resonant when they are near a quarter-wavelength. When more and more loops are added in parallel, the assembly eventually becomes a closed resonant box as shown at E. This box, called a *resonant cavity*, is a quarter-wave in radius or a half-wave in diameter.

A resonant cavity displays the same resonant characteristics as a tuned circuit composed of a coil and capacitor. In it there are a large number of current paths. This means that the resistance of the box to current flow is very low and that the  $Q$  of the resonant circuit is very high. While it is difficult to attain a  $Q$  of several hundred in a coil of wire, it is fairly easy to construct a resonant cavity with a  $Q$  of many thousands. Although a cavity is as efficient at low as at high frequencies, the large size required at low frequencies prohibits its use.

For example, at one megacycle a resonant cavity would be a cylinder about 500 feet in diameter. When the frequency is in the vicinity of 10,000 megacycles, the diameter of the cavity is only 0.6 inch. This makes the cavity smaller than a conventional tuned circuit. Therefore, equipment which operates at a frequency of 3000

mc or above usually uses resonant cavities as resonant circuits.

If there were no losses in the cavity, the waves would bounce back and forth between the end walls indefinitely. However, just as in any resonant circuit, some energy is lost during each cycle. Usually, a source of microwave energy feeds the cavity continuously so as to overcome the losses.

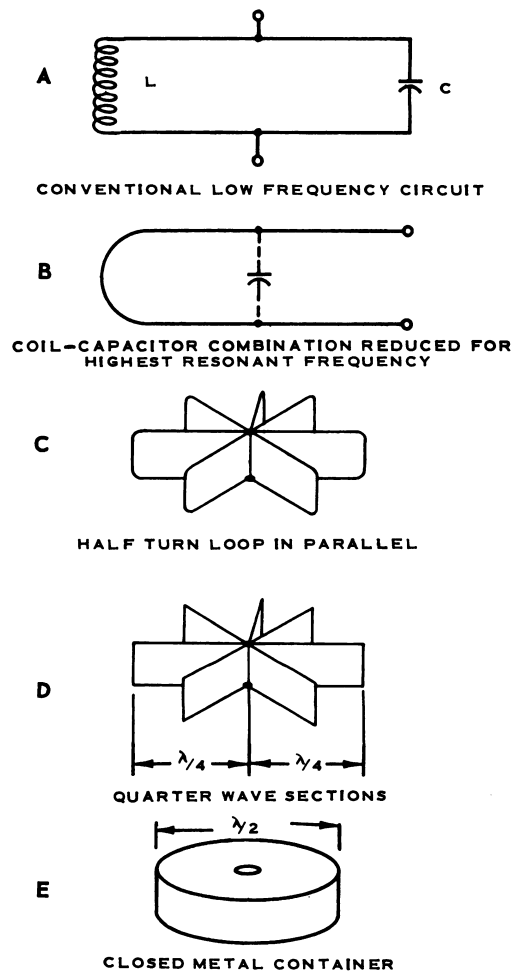


Figure 11-48. Development of Cavity from  $\lambda/4$  Sections

### The Fields in a Cavity

A resonant cavity may be compared to a waveguide, since its operation is best described in terms of the fields rather than in terms of the currents and voltages present. As in waveguides, the different field configurations in cavities are called *modes*. In Figure 11-49 which shows the dominant mode of the cylindrical cavity, note



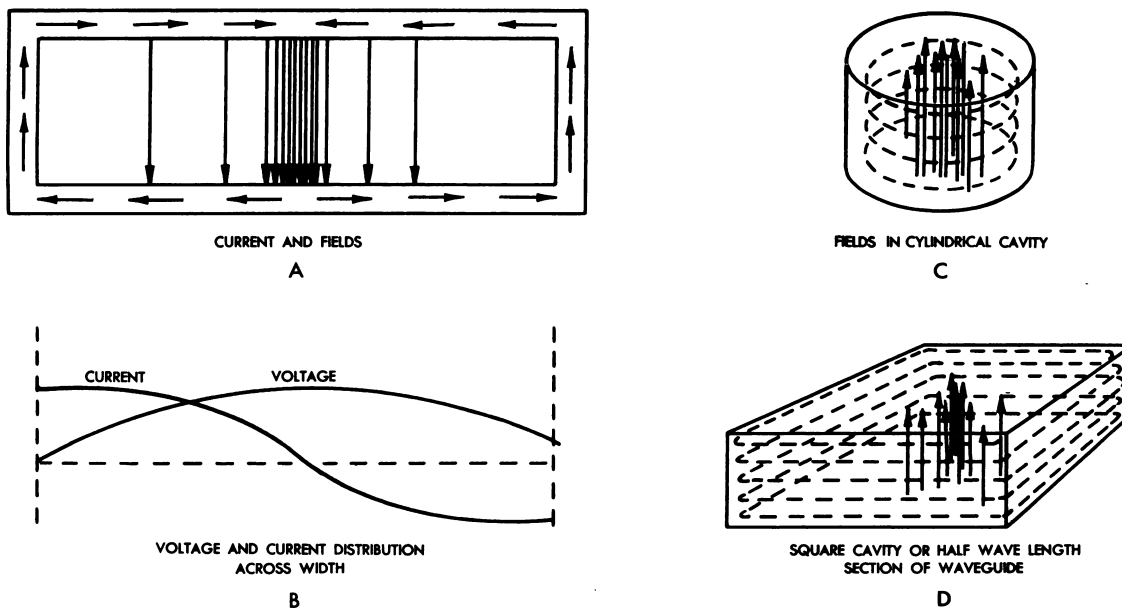


Figure 11-49. Voltages, Currents and Associated Fields for Simple Mode in a Cavity Resonator

that in A the voltage is represented by E-lines between the top and the bottom of the cavity. The current flows in a thin layer on the surface of the cavity, due to skin effect. The strength of the current is indicated by graduated arrows.

The magnetic field is strong where the current is high. The strongest H-field is at the vertical walls of the cylinder and diminishes toward the center where the current is zero. This is due to the standing waves on the quarter-wave section. The E-field is maximum at the center and decreases to zero at the edge, where the vertical wall is a short circuit to the voltage. The curves of E-field and H-field density are shown in Figure 11-49B. Figures C and D show two types of cavities with their fields represented.

The modes in a cavity are identified by the same numbering system that is used with waveguides, except that a third subscript is used to indicate the number of patterns of the transverse field along the axis of the cavity (perpendicular to the transverse field). For example, the cylindrical cavity shown in Figure 11-49C is a form of circular waveguide. The axis is the center of the circle. The transverse field is the magnetic field. Therefore, it is marked TM. Around the circumference there is a constant magnetic field. (The H-lines are parallel to the circumference.) Therefore, the first subscript is

zero. The distance across the diameter is one half-wave. Thus the second subscript is one. Through the center along the axis the H-field strength is a constant zero. This makes the third subscript zero. Therefore, the complete description of the mode is  $TM_{0,1,0}$ .

When a half-wave section of waveguide is closed on both ends in the form of a rectangular cavity, as shown in Figure 11-49D, standing waves are set up and resonance occurs. The simple mode in this cavity is the same as the dominant mode of a rectangular waveguide, that is,  $TE_{0,1}$ . The third subscript of the mode, which is determined by the plane of the E-field, is 1. Thus, the complete description of the simple mode in the rectangular cavity at D is  $TE_{0,1,1}$ .

Cavities may have various physical shapes, for any chamber enclosed in conducting walls resonates at several frequencies and produces a number of modes. Note in Figure 11-50 examples of several types of cavities. The Q of each cavity is indicated. Of those shown, the cylindrical cavity is useful in wavemeters or in frequency-measuring devices. The cylindrical ring type is used in super-high-frequency oscillators as the frequency determining element. The section of waveguide, which is shown diagrammatically, is used in some radar systems as a mixing chamber for combining signals from two sources.

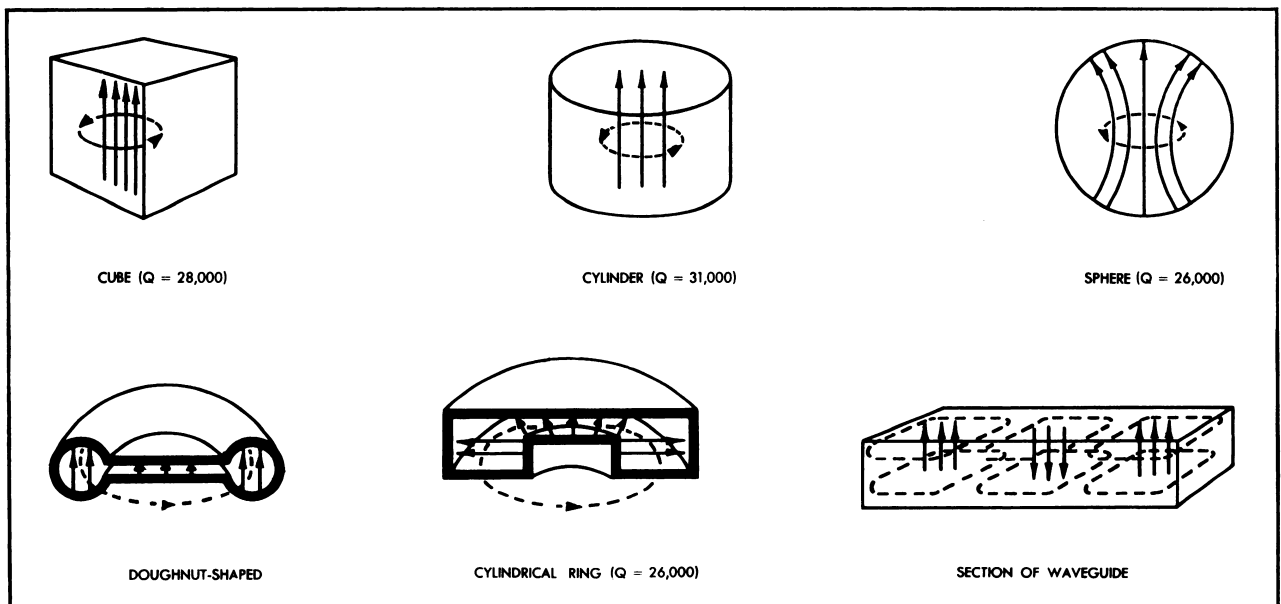


Figure 11-50. Several Types of Cavities

### Exciting the Cavity

There are three principal ways in which energy can be inserted into, and removed from, a cavity resonator, Figure 11-51. The first is by inserting a probe into the cavity. The current flowing in the probe sets up E-lines parallel to it and they, in turn, start oscillations. Another method uses a magnetic loop. The loop is placed in the region where the magnetic field will be located. The currents in the loop start an H-

field in the cavity. Either of these methods can be used to remove energy from, or to put energy into, the cavity.

A third method uses a cylindrical ring-type cavity. In this method the energy is placed into the cavity by clouds of electrons which are virtually shot through the holes in the center of a perforated plate. As each cloud goes through, it creates a disturbance in the space inside the cavity until a field is set up. In terms of current,

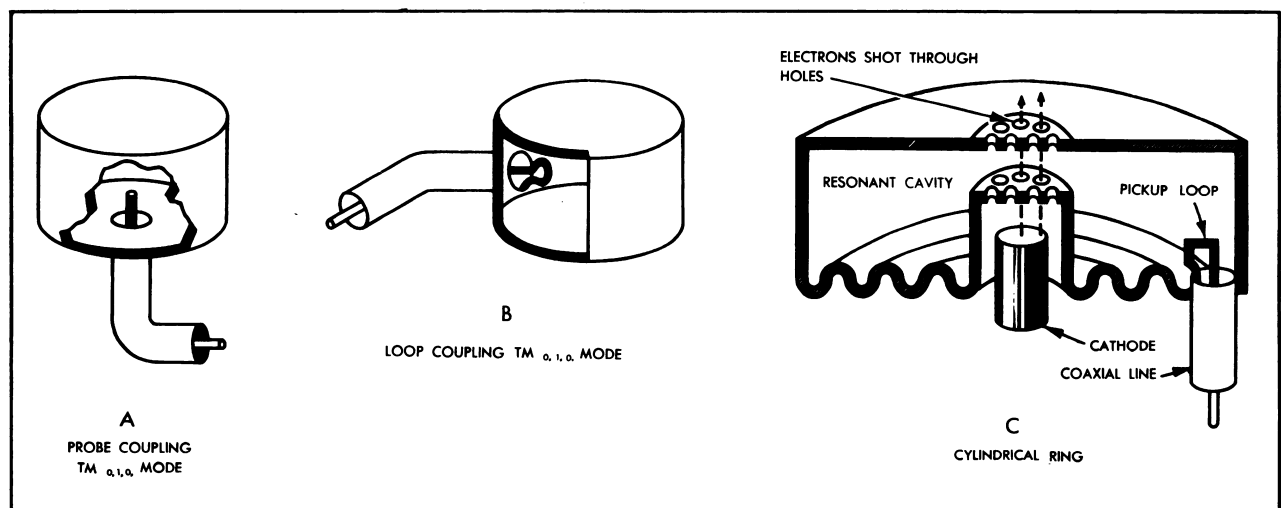


Figure 11-51. Methods of Exciting the Cavity

the approaching cloud of electrons makes the perforated plate positive by repelling the electrons away from it. This current starts an H-field. Energy may be removed from the cavity by placing a loop at the outside edge.

Other methods of feeding the cavity are discussed later in the analysis of systems using cavities.

### Varying the Resonant Frequency of the Cavity

Three methods for setting the resonant frequency in a cavity are shown in Figure 11-52. One method uses a cylindrical cavity with an adjustable disc. When a  $TE_{0,1,1}$  mode is used, the size of the cylinder may be changed along the axis to change the resonant frequency. The smaller the volume of the cavity, the higher the resonant frequency. The movement of the disc may be calibrated in terms of frequency. In the case of high-frequency equipment, a micrometer scale is usually used to indicate the position of the disc, and a calibration chart is used to determine the frequency.

A second method for varying the frequency uses threaded plugs that are inserted in the cavity. The plug reduces the strength of the magnetic field in the cavity in a manner similar to reducing the inductance of the tuned circuit. The deeper the plug extends into the cavity, the higher the frequency.

In a third method the interior of the cavity, which is part of the interior of a vacuum tube, is sealed and evacuated. In this method a fre-

quency change occurs whenever the top and bottom of the cavity are moved away from, or toward, each other. This is accomplished by turning a screw at the end of a lever. As the screw is turned, the two pieces of metal to which the lever is attached are either pulled together or allowed to spring apart. This moves the top of the cavity up and down with respect to the bottom. As the distance varies, the volume and the capacity between the top and bottom of the cavity are changed. As the change in capacity is the chief result of the change in the distance between the plates, the resonant frequency is inversely proportional to the distance from the top to the bottom.

Another way of changing the frequency is by changing the method of exciting the circuit. This can be done by tuning the exciting loop either capacitively or inductively. This can occur either accidentally, due to improperly tuned circuits, or deliberately, as a means of tuning the cavity.

### Uses of Cavities

There are many ways in which resonant cavities are used in radar. One important use is as a tuned circuit in 10,000-mc oscillator circuits. For example, the klystron shown in Figure 11-53A, which can be used as the local oscillator in a radar receiver, is tuned by changing the resonant cavity. In the klystron the cathode emits electrons that pass through the grids toward the plate. When they pass through this

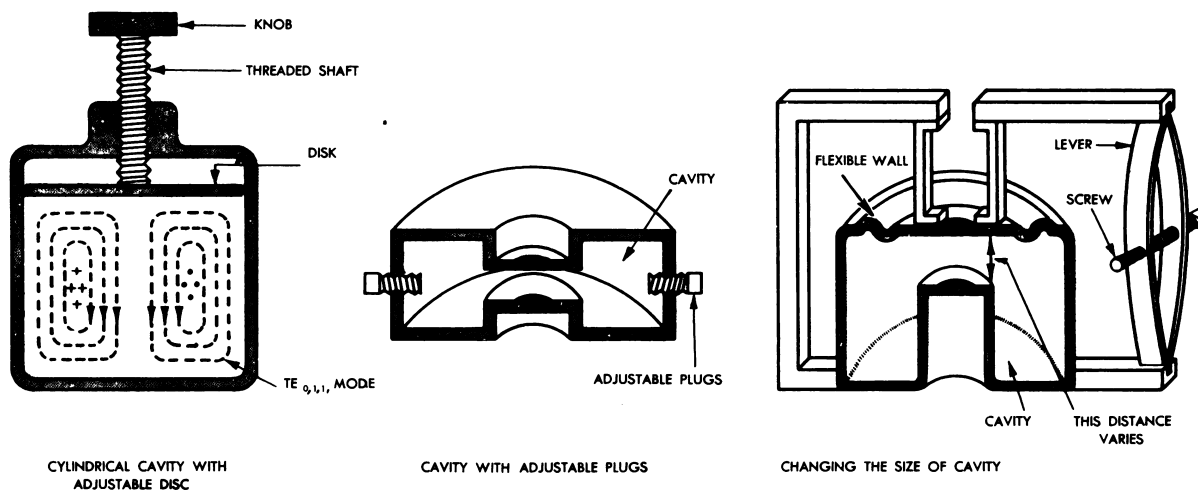


Figure 11-52. Method of Changing the Frequency of a Cavity

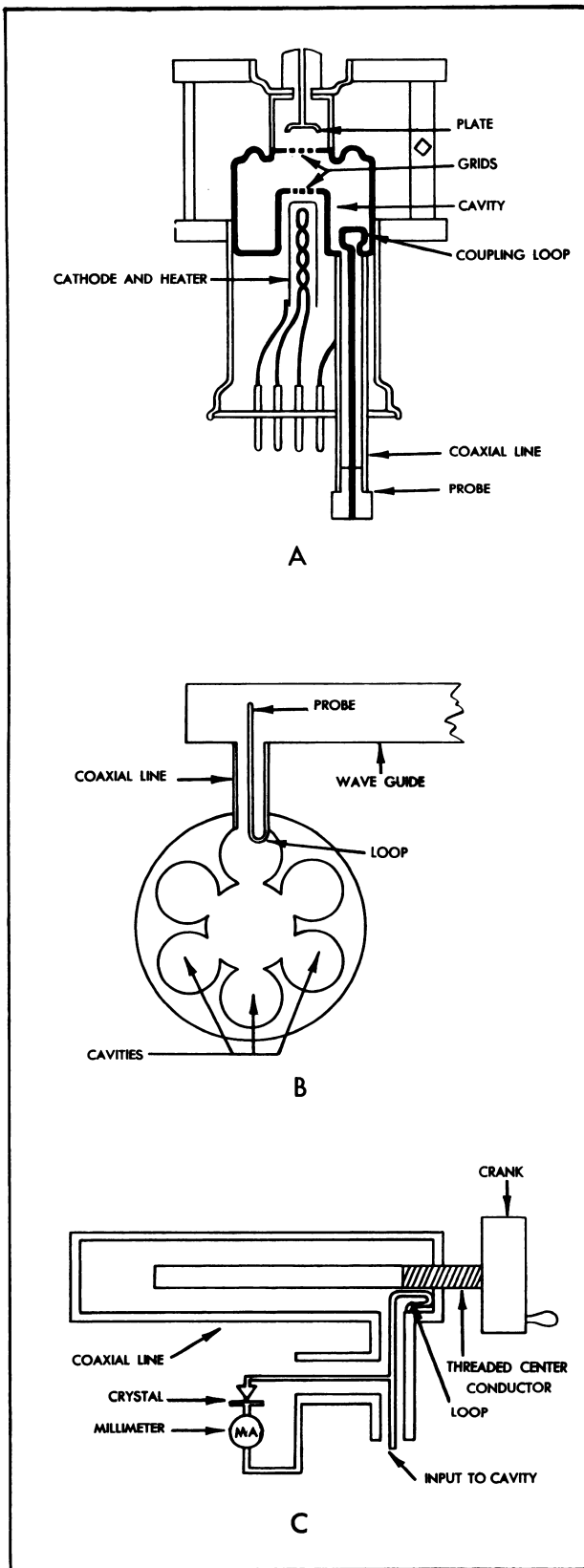


Figure 11-53. Uses of Cavities

area, they disturb the field and excite the cavity. The resonant frequency is changed by the spacing of the grids.

Another use of resonant cavities is the series of cavities arranged around a circle in the magnetron oscillator as shown in Figure 11-53B. These cavities, which are coupled one to another through capacitance at the openings, are excited by moving electrons. The energy from the electrons is passed around the ring from all cavities to the one with a loop. This one serves to transfer the energy to the waveguide. From the waveguide, the energy goes to the antenna where it is radiated into space.

Since the frequency of the transmitter is fixed, there are no frequency adjusting devices on a waveguide. However, the different speeds at which the electrons travel and the variable impedances in the waveguide cause either a change in frequency, or oscillations in a mode which differs from the one intended.

Cavities also may be used as wavemeters — devices for measuring frequency. One type of wavemeter consists of a cylindrical cavity with an adjustable disc. Another type operates essentially as the first one but uses a coaxial cavity and operates in a different mode, as shown in Figure 11-53C. In the second type the signal is introduced into the cavity by a loop located in one end of the cavity. The signal is strongest when the loop is at a high current point in the standing wave.

The standing wave of current is maximum at odd multiples of quarter-wave distances from the open end of the guide. When the length of the threaded center conductor is varied by the crank, the distance from the open end to the loop can be made equal to an odd multiple of a quarter-wave. This causes the current maximum to occur at the loop, and this, in turn, causes the input impedance at the loop to be zero. When the distance from the open end of the guide to the short (closed end) is a quarter-wave or any odd multiple of a quarter-wave, the current introduced into the guide will be at the location of parallel resonance.

The threaded shaft shown in Figure 11-53C can be calibrated either in wavelength or in frequency. The input impedance can be indicated by any suitable indicator, such as a crystal rectifier and a DC milliammeter. In operation, when the input impedance is zero, the signal is shorted

out. In this case the meter will read zero. To check the frequency for a zero impedance reading, turn the crank until a dip occurs on the meter and read the frequency on the threaded shaft.

A cavity can be used as a mixing chamber. When two or more signals are put into a cavity, it is possible to remove the combination of their signal voltages. An example of this use is discussed later in this chapter.

Another use of a cavity is as an impedance-matching device such as shown in Figure 11-54; this illustrates a small waveguide connected to a large waveguide. Normally, when you connect the two waveguides together, there will be reflections in the waveguides. However, a small

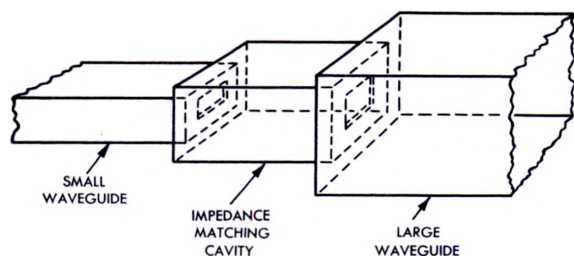


Figure 11-54. Cavity Used for Matching an Impedance Device to One of Different Impedance

section of waveguide or an intermediate size can be matched without reflections, provided that plates are used at the junction to cancel the reactive effects resulting from the different sizes of the waveguides. Thus, although the cavity itself has standing waves in it, there are none on either waveguide.

Another use of a cavity is a *ringing circuit* in a resonant circuit. It is called ringing circuit because it oscillates for many cycles after being started by an external circuit. In this use the cavity is resonant; once started, it will continue oscillating for a few microseconds after the source of voltage has been removed. This action is analogous to a bell that continues its tonal vibrations for many seconds after being struck.

In the 3-cm (10,000-mc) ringing circuit shown in Figure 11-55, the cavity itself is a cylinder which is 3 inches in diameter and 10 inches long. A movable piston in the cylinder determines the volume of the cavity. Because the position of the piston is calibrated in megacycles, it serves as a frequency meter. The cavity is energized by a signal that enters from the small waveguide through a small aperture. The energy is put into the short waveguide by a probe which, in turn, is fed by a coaxial cable. A good method to put the signal into the coaxial cable is to connect a small antenna to the end of the cable.

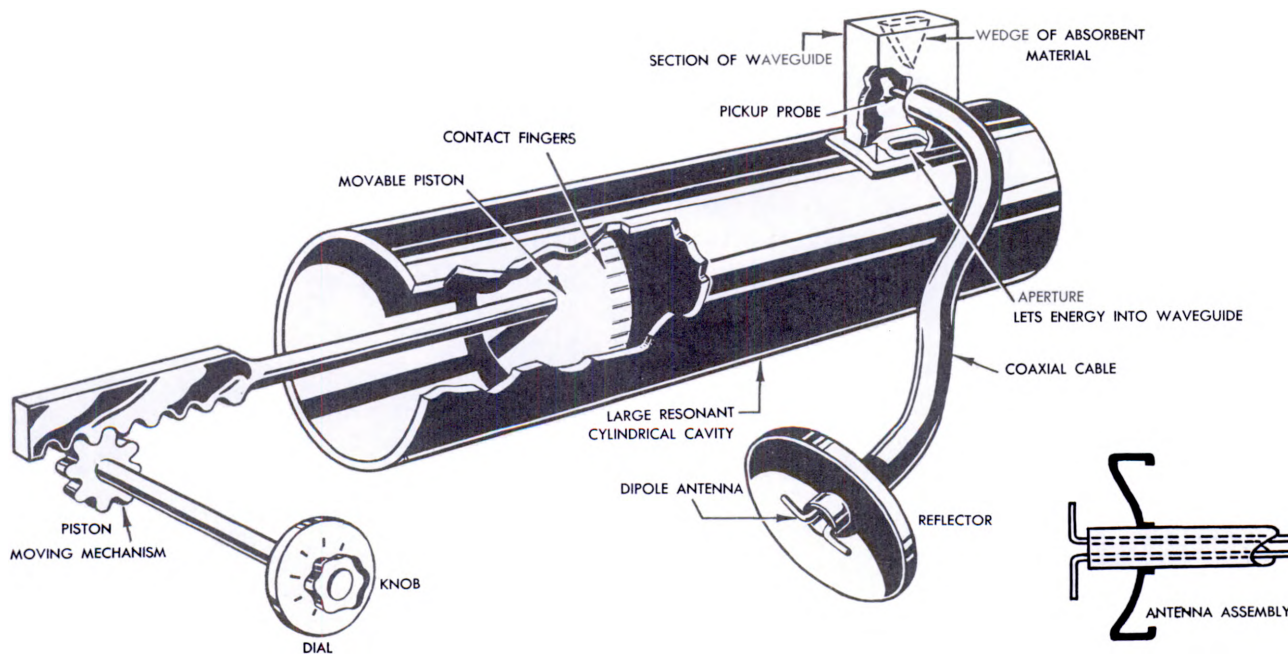


Figure 11-55. 3-cm Ringing Circuit



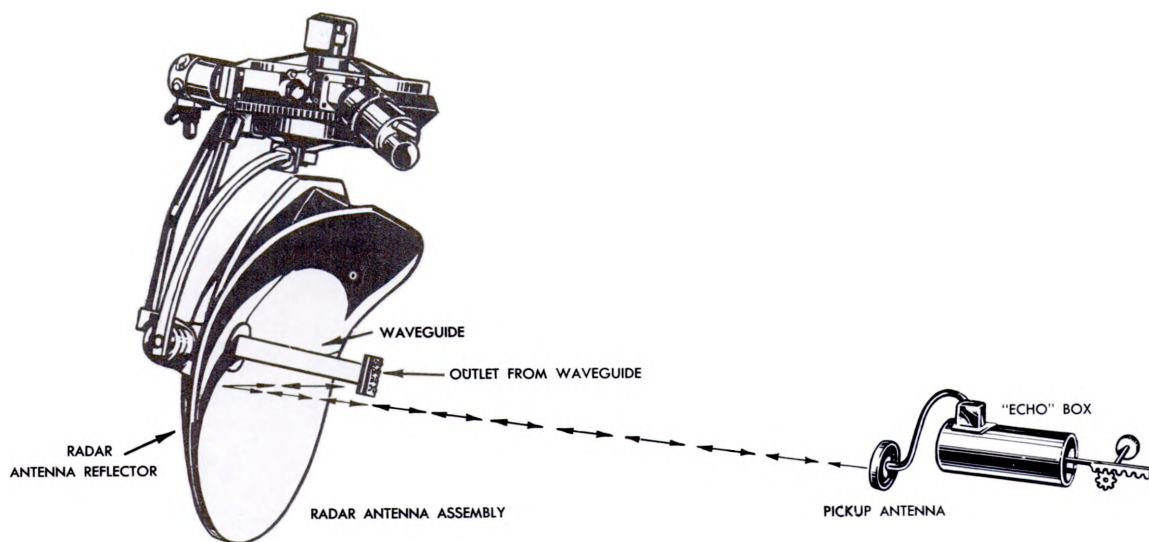


Figure 11-56. Cavity as an Echo Box

The resonant cavity shown in Figure 11-55 is used as an echo box as illustrated in Figure 11-56. Echo boxes are resonant cavities having a very high  $Q$ . They are used with microwave radar sets to provide an artificial or "phantom" target, which may be used to tune the receiver to the transmitter when no real targets are available.

This is how the echo box works. When the radar transmitter is turned on, energy is fed into the echo box cavity. There it sets up violent oscillations, and the cavity itself acts as a transmitter for a period of 20 microseconds after the pulse generated by the radar transmitter ends. (When you view the oscillations on an oscilloscope, their waveshapes appear as shown in Figure 11-57.) The energy which the echo box radiates is picked up by the small pickup antenna

located in front of the radar antenna assembly and is fed into the radar receiver. The receiver displays it on the indicator screen. Since this signal simulates a radar echo, the cavity which transmitted it is appropriately called an echo box. It is thus a means of supplying a signal for tuning the receiver when there is no real target available.

## MICROWAVE FERRITES

### General

Recent developments in microwave techniques have produced a series of waveguide fittings that use the unusual properties of a new type of magnetic material. These materials are called ferrites; they are ceramic materials containing iron and have unusual magnetic and electrical properties. Ferrites have wide use in waveguide applications. They are used as isolators between transmitter and antenna, as waveguide switches, as unidirectional attenuators, and as commutators (Faraday rotation).

Ferrites are compounds of iron, zinc, manganese, magnesium, cobalt, aluminum and nickel oxides. They are manufactured by pressing into shape the required mixture of the finely divided metallic oxide powders and then firing the shaped mixture at about 2000°F. The product is a ceramic with high electrical resistance.

Ferrites have been used commercially for several years. For instance, there are nickel-zinc

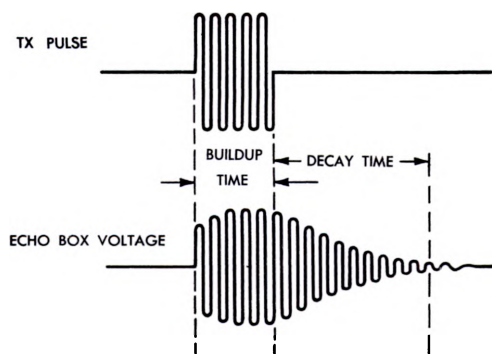


Figure 11-57. Oscillations in Echo Box



ferrites in the deflection coils of flyback transformers in television sets; a powdered ferrite is used in magnetic recording tape. Ferrites behave as iron alloys at low frequencies, but at high frequencies their high electrical resistance prevents eddy currents and resonance takes place within the iron atoms themselves. These unusual effects make it possible to use ferrites for special applications in microwave circuits.

### Fundamental Theory of Ferrites

A fundamental property of atoms is that both electrons and protons spin on their own axes, which was explained in Chapter 2. In addition, of course, the electron revolves around the nucleus. An analogy is the solar system, where the earth rotates on its axis as it revolves around the sun. As the electron spins, it creates a magnetic moment, or field, along its spin axis. This spinning charge appears as a current flowing around a loop. The atoms with more electron spins in one direction than another act as small magnets. The mutual action of all these atoms explains the magnetic properties of magnetic materials.

If a spinning electron is placed in a static magnetic field, the electron's magnetic moment becomes aligned with the static field. The magnetic moment and its alignment with a DC magnetic field is shown in Figure 11-58.

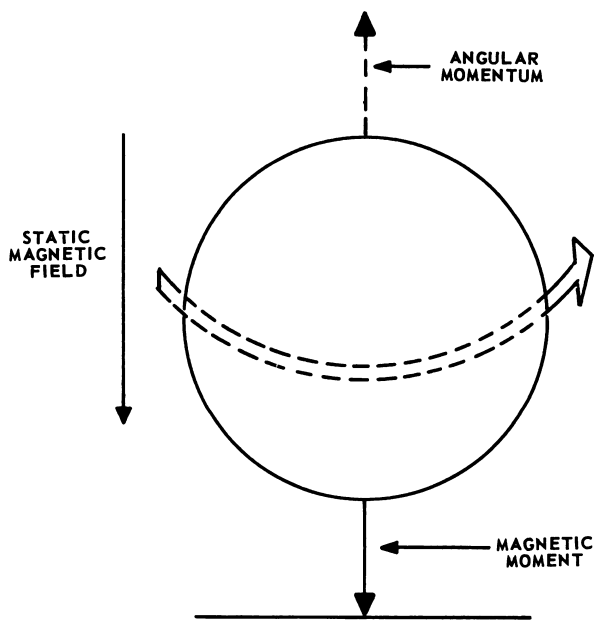


Figure 11-58. Electrons in DC Magnetic Field

### Gyroscopic Action

Electrons, because of their spinning motion, behave like very small gyroscopes. When a force is applied to the spin axis of an electron that would cause it to tilt, the electron will behave like any other gyroscope. It will precess, or wobble. Precession is defined as a movement of the axis of rotation at right angles to its original axis. Figure 11-59 shows a gyroscope mounted to a stick so that the stick can pivot freely. Even with the gyroscope spinning, the stick will hang straight down, due to gravity. However, if you try to move the stick from side to side, the gyroscope will force the stick to move around in a circle, or precess.

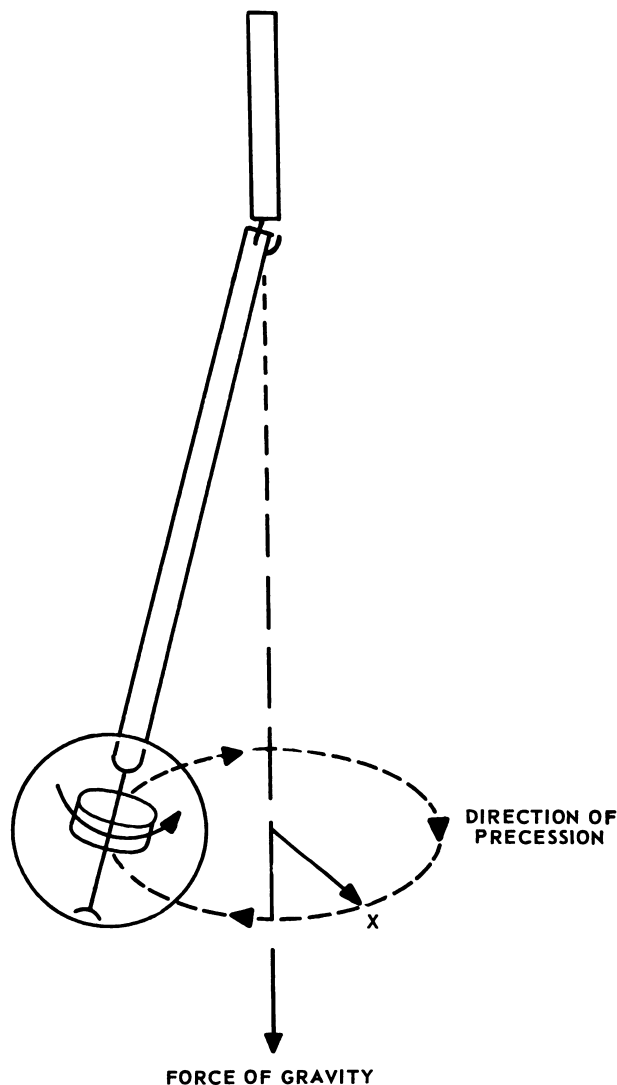


Figure 11-59. Precession of a Gyroscope

The direction is determined by the direction of rotation of the gyro rotor, and the frequency by the gravitational force and the momentum of the gyroscope. This is shown in Figure 11-59. The natural precession frequency could be increased by increasing the force of gravity. A rotational force applied to the stick at the natural precession frequency will displace it from the vertical by a large amount. (The precession path shown in Figure 11-59 will have a greater diameter.) Thus, there will be a large amplitude of moment. A rotational force applied at any other frequency will produce a much lower displacement. This is similar to the feedback in an oscillator. With feedback at the right frequency, the amplitude of oscillation is larger than when the feedback is off frequency.

### Electron Action

Electrons behave much like gyroscopes, but gravity has little effect on them. Instead, a steady magnetic field is applied to line up the

axes of the spinning electrons. This field causes any precession to die out quickly. Now when an alternating field is applied at right angles to the DC field, the electrons precess, or wobble, just the way the gyroscope and stick did when a sideways force was applied.

The natural precession frequency of an electron in a DC field is between 3000 and 9000 megacycles, depending on the field strength. If we apply an AC magnetic field at the natural frequency, the precessional motion will build up. This increases the frictional damping effects because the entire iron atom is vibrating. The ferrites dissipate as heat the energy extracted from the AC field.

### Applications

**ATTENUATOR.** One application of ferrites is as an attenuator. Figure 11-60 shows a piece of ferrite placed in the center of a waveguide; a steady magnetic field is applied as shown. This arrangement will attenuate frequencies at the

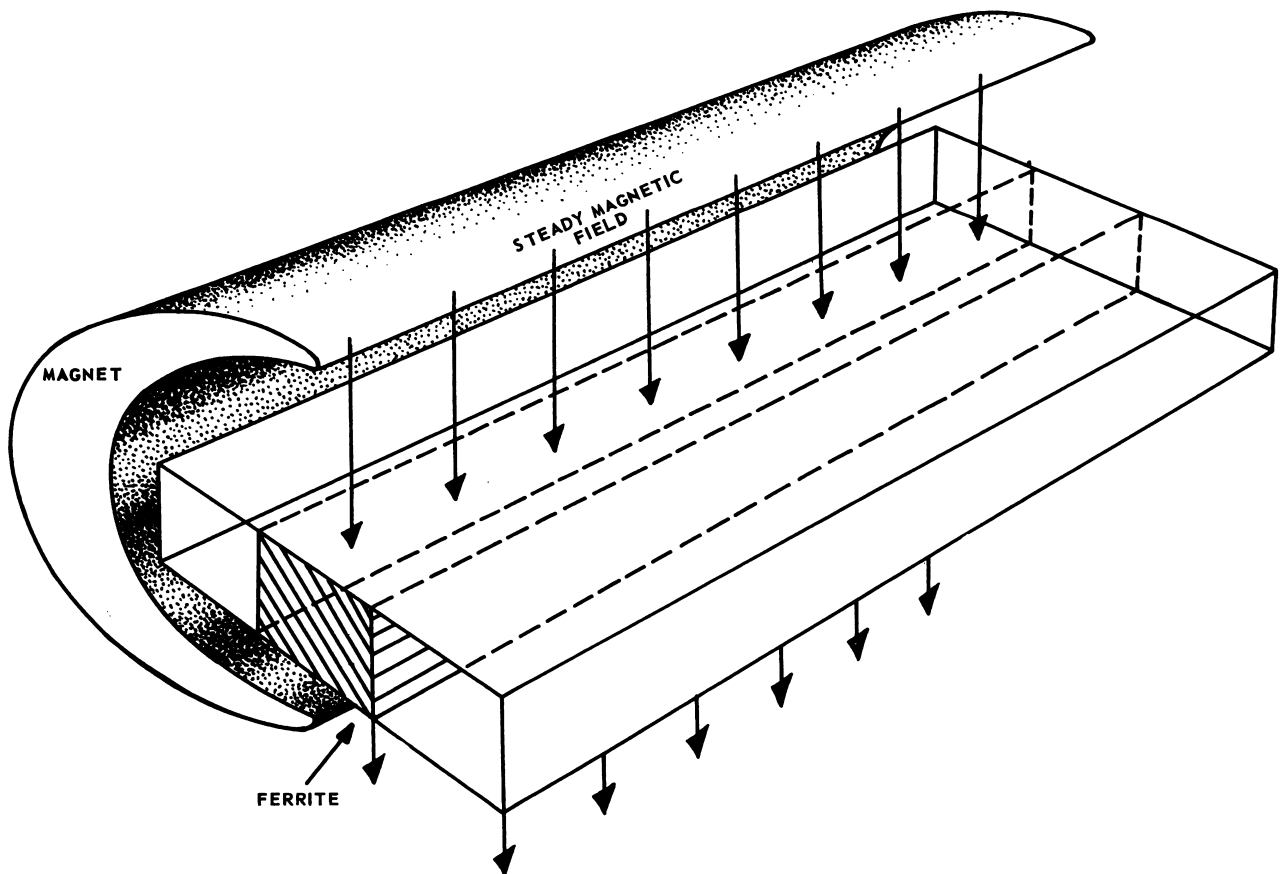


Figure 11-60. Ferrite Slab Mounted in Waveguide

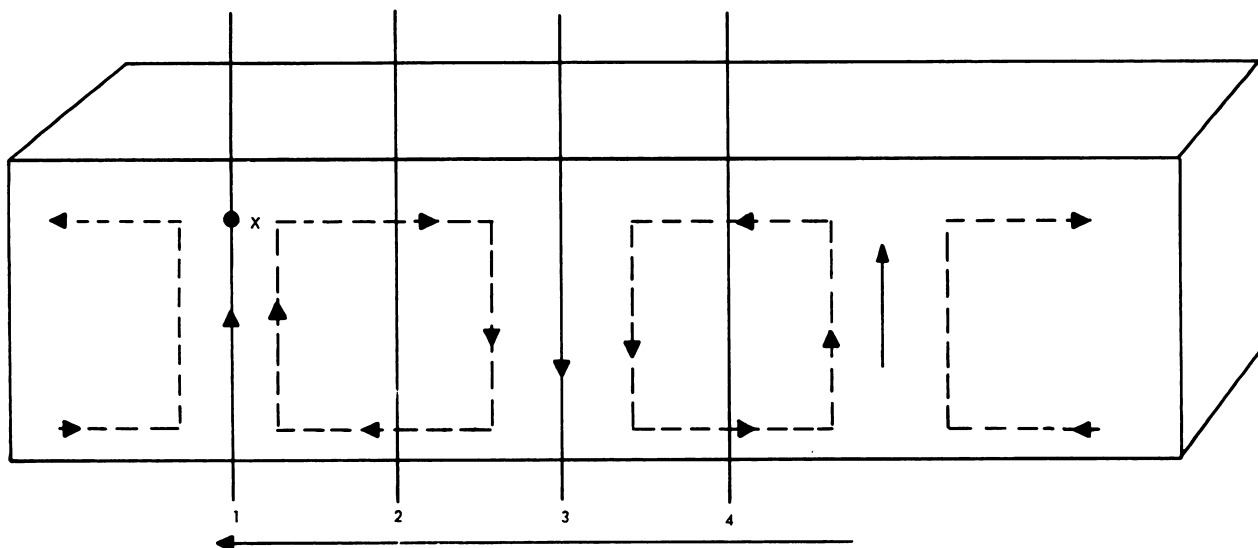


Figure 11-61. Effective Clockwise Rotation of a Magnetic Field

resonant frequency of the electrons in the ferrite, whereas other frequencies will be attenuated very slightly. Changing the strength of the DC field produces a change in the frequency that will be attenuated, although this occurs over only a limited range.

The DC field is produced by current flowing through a coil wound around the waveguide. The strength of the field, which depends upon the current flowing through the coil, determines the frequency of precession. Usually the ferrite attenuator is in the form of an adjustable vane of ferrite extending into the waveguide. The farther the vane extends into the waveguide, the greater the attenuation, since more of the r-f energy must travel through the ferrite.

**ISOLATORS.** Another application of ferrites is that of an isolator. When used as an isolator, the ferrite allows energy to travel in one direction, but absorbs energy traveling in the opposite direction. Figure 11-61 depicts an electromagnetic wave traveling from right to left. It illustrates how the wave, at a point off the center line of the guide, will appear as a rotating magnetic field. At one instant shown in Figure 11-61, the magnetic field at Point X is pointed up. When the magnetic field at Point 2 reaches Point X, the magnetic field will be directed to the right. When Point 3 reaches X, the magnetic field is downward, and when Point 4 on the wave arrives at X, the field is directed to the left.

Thus, as the wave passes Point X, the magnetic field appears to rotate in a clockwise direction. At any point off the center of the waveguide, the magnetic field will appear to rotate as the electromagnetic wave passes. This same analysis can be used to show that with a wave traveling from left to right the magnetic field appears to rotate counterclockwise at Point X.

Now let us place a section of ferrite in the waveguide at X. This is shown in Figure 11-62, which illustrates a simple isolator consisting of a piece of waveguide, a permanent magnet and a section of ferrite. The ferrite's electron resonant frequency and the microwave frequency are made the same by changing either the magnetic field strength or the microwave frequency. When the frequencies are the same, a wave traveling from left to right in the waveguide produces a rotating force in the direction of the natural precession of the electrons in the ferrite. The amplitude of precession increases, taking power from the electromagnetic wave. This power is dissipated as heat in the ferrite.

A wave that is traveling from right to left in this waveguide acts as a rotating force on the electrons to oppose the natural precession. This will not increase the amplitude of the precession and energy is not absorbed from the electromagnetic field. About 0.4 db attenuation takes place in a wave traveling from right to left, but as much as 10 db occurs in a wave traveling from left to right.

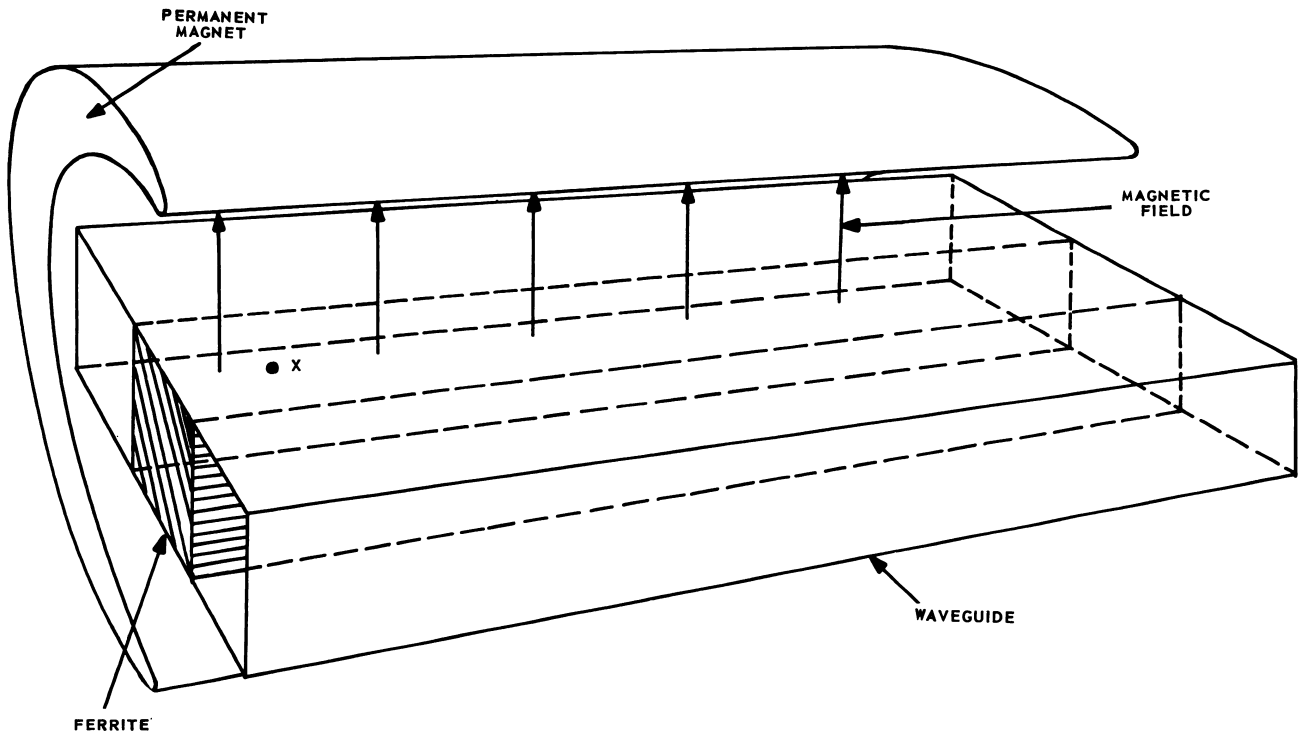


Figure 11-62. Simple Isolator

**FARADAY ROTATION.** Another effect takes place when microwaves are passed through a piece of ferrite in a magnetic field. The plane of polarization of the wave is rotated if the frequency of the microwave is above the resonant frequency of the ferrite electrons. This is known as the Faraday rotation effect. When r-f energy enters the ferrite material, the magnetic moment of the electron precesses as usual but at a different frequency from the r-f. The H-lines within the ferrite now are the resultant produced by vectorial addition of the rotating magnetic moment and the r-f field. A new r-f field, which is rotated from the original r-f field, results. The amount of rotation is determined by the DC magnetic field and the length of the ferrite.

Figure 11-63 shows a ferrite rod that is placed lengthwise in the waveguide. The DC magnetic field is set up by a coil. Now assume that a wave that is vertically polarized enters the left end of the waveguide. As it enters the ferrite section, it will set up limited precession motion of the electrons. The interaction between the magnetic fields of the wave and the precessing electrons rotates the polarization of the wave. With the correct dimensions of the ferrite rod,

the wave is polarized at a  $45^\circ$  angle from the original. Different dimensions of the rod and magnetic field strengths produce other shifts in polarization.

**DUPLEXER AND DIPLEXER.** The duplexer is another application of the ferrite. As you know, a duplexer is used in radar to allow one antenna to both transmit and receive. The ferrite duplexer consists of a folded hybrid, a phase-shift section (ferrite), a short slot hybrid and two output arms. Power from the transmitter is fed into the folded hybrid. Here it is split into two parts of equal size and with the same phase relationship. These two, however, will be  $180^\circ$  out of phase to the receiver and will cancel.

After splitting into two parts, the transmitter energy goes to the ferrite section in phase. Here one channel, let's say the lower, shifts the energy  $90^\circ$  in phase, while the energy in the other channel (upper) passes through unchanged. Next, the energy goes to the short slot hybrid. Here the energy in the lower channel passes through unchanged, while that in the upper channel is shifted  $90^\circ$ . Now the energy in the two channels is back in phase and is radiated by the antenna.

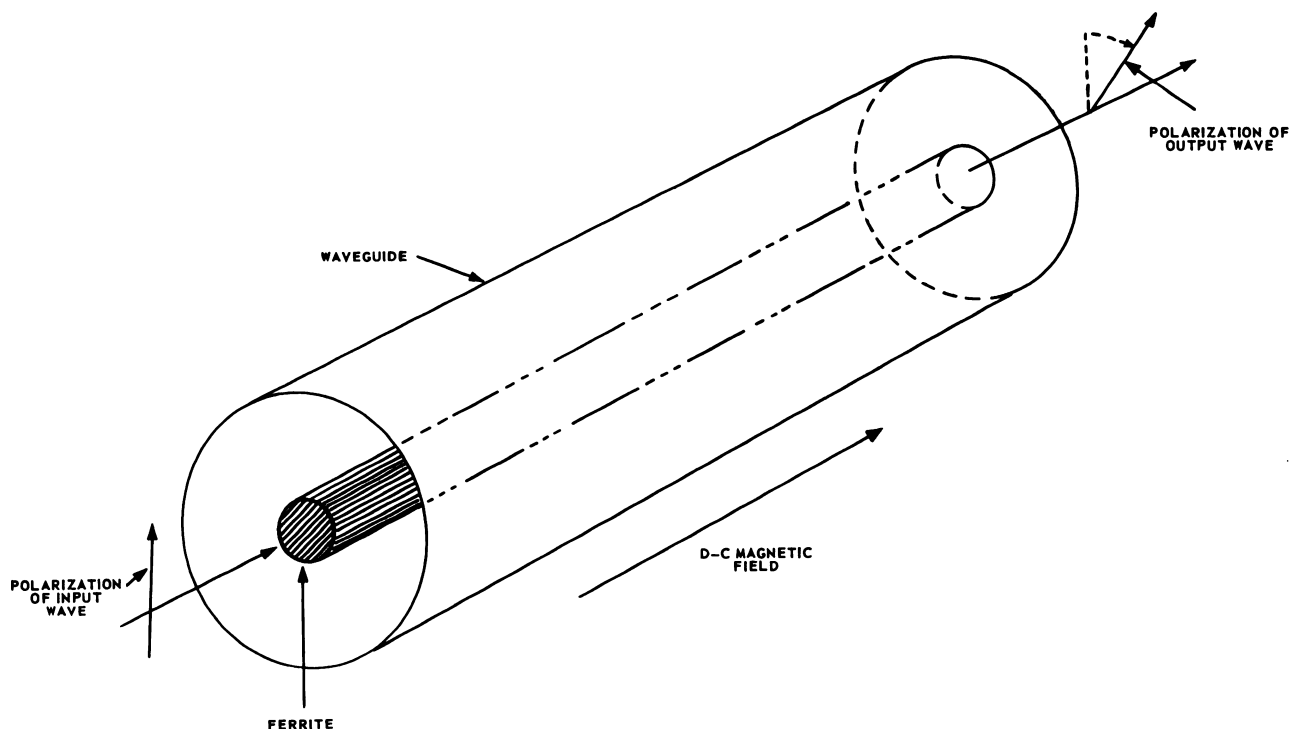


Figure 11-63. Faraday Rotation

When energy is received by the antenna it is divided into two channels when it enters the short slot hybrid. The energy in the lower channel passes through the ferrite section and enters the folded hybrid still in the same phase. There is no phase shift. But the energy in the upper channel is shifted  $90^\circ$  in the short slot hybrid and another  $90^\circ$  in the same direction in the ferrite section. The energy in the upper channel has been shifted  $180^\circ$ . Since the energy in the two channels in the folded hybrid is  $180^\circ$  out of phase, the received energy enters the receiver waveguide section. However, the received energy would leave the folded hybrid for the transmitter section  $180^\circ$  out of phase and here would cancel. This prevents any energy from being lost in the transmitter section.

A *diplexer* is used in radar to allow two different transmitter frequencies to be fed into the same waveguide without interaction between the two transmitters. The diplexer section itself consists of two resonant cavities, which are tunable to different frequencies, feeding a single waveguide section. The two different transmitter channels each feed into one of the resonant cavities. The energy from one resonant cavity can-

not feed back into the other transmitter channel, due to the resonant cavity being tuned to the other frequency. The diplexer prevents feedback from one transmitter to the other. The energy from the diplexer is fed into the folded hybrid section of the diplexer discussed in the preceding paragraphs.

#### Ferrite Cores

The ferrite core is used in the memory units of computers. In this application the magnetizing characteristic of the ferrite is used. The electrons in the circular core of the ferrite material align themselves with a magnetic field. The magnetic field can be produced by current passing through wire looped around the core or passed through the center of the core. The direction of current determines the direction of magnetization.

Ferrites in microwave applications are becoming familiar devices. Typical applications include: load isolation between a microwave source and its load to prevent reflections from affecting the generator; modulation of microwaves with pure amplitude modulation; and fast action radar duplexers.

## A COMPLETE WAVEGUIDE R-F SYSTEM

In a complete waveguide r-f system the signal source is the transmitter unit. The antenna that transmits the output is located as near as possible to the radar transmitter. A principal requirement in an efficient radar system is that energy must be transferred from the transmitter to the antenna with minimum losses, for any losses that occur will shorten the maximum range and thereby decrease the efficiency of the entire radar system. Since the transmitting antenna is pointed at the *target* at the time the transmitter pulse leaves, and since it is conveniently located, it is preferable to use it as the receiving antenna, too. Similarly, it is possible to use the same waveguide system for transferring the received energy to the radar receiver.

When the same elements are used for both transmitting and receiving, it is necessary to use a high-speed automatic switch to prevent the powerful transmitter pulse from going into the receiver circuit and causing serious damage. The TR and ATR boxes discussed earlier perform this function in many radar sets.

The highly directional antenna is made rotatable in order to cover all areas. To make rotation possible, signals are coupled to the antenna through vertical and horizontal rotating joints.

Since airborne equipment is flown at high altitudes, temperature changes will cause moisture condensation inside the waveguide. As water causes extremely high losses, the interior of the waveguide must be maintained at a higher pressure than the outside to drive the moisture out and keep it out. Furthermore, all joints have to be airtight to maintain this pressure.

Figure 11-64 shows a waveguide system that uses the components just discussed. The source of the r-f energy is the magnetron tube which is shown at the right in the lower part of Figure 11-64. A probe couples it to the waveguide, which operates in the  $TE_{0,2}$  mode. A very short section of waveguide is fixed permanently to the magnetron output. When the magnetron is installed, its waveguide section is connected to the end of the main waveguide. In operating conditions the transmitter energy travels down the main waveguide where it enters a section which has two T-junctions — one called the ATR box, and the other the TR box and mixing chamber.

In each T-junction, there is a spark gap at a distance of a quarter-wave from the center of the main guide. Both these gaps arc over when the transmitter is on. Thus, no energy goes past either spark gap tube into the T-junction waveguides. The energy flow is continuous down the main guide to the antenna. Each section is connected to the next with a choke joint.

In going to the antenna, the signal passes through a special section that contains a pair of probes spaced a quarter-wave apart. Test equipment can be inserted at each probe to measure the signal strength at the probe. The ratio of the two signal strengths is the standing-wave ratio. In going through the stationary transmitter unit to the moving antenna, the signal goes through a flexible section of waveguide that insulates the transmitter unit from the mechanical vibration of the antenna.

The antenna is designed to rotate horizontally and to tilt vertically. Thus, the signal is transferred to a rotating coaxial joint twice before being radiated, and each time it is immediately transferred back to a waveguide again. The antenna itself is a resonant cavity on which there are a pair of slots one half-wave apart. As you can see in the enlargement, the waveguide is tapered in the narrow dimension to fit between the slots. The effect of the cavity is as though the waveguide were split in half and each side bent through  $180^\circ$ . Tapering the waveguide prevents the reflections that would occur if the size were changed abruptly. Tapering in the narrow dimension does not affect the operation of the waveguide, since this dimension is not critical. The actual impedance match is accomplished by adjusting the matching screw. This adjustment is made at the factory and soldered in place.

The energy comes out of the back of the cavity and goes to the large reflector. The reflector sends the energy out forward in a narrow beam. The energy which returns from the radar target is reflected by the reflector into the slots. It passes through the slots into the waveguide and travels toward the magnetron, where it stops at the ATR box. The length of the T-junction is a half-wave from the center of the waveguide, so the closed end of the section reflects a short circuit at the outer center of the main waveguide. This reflects the signal back along the guide to the mixing chamber. Of course, neither



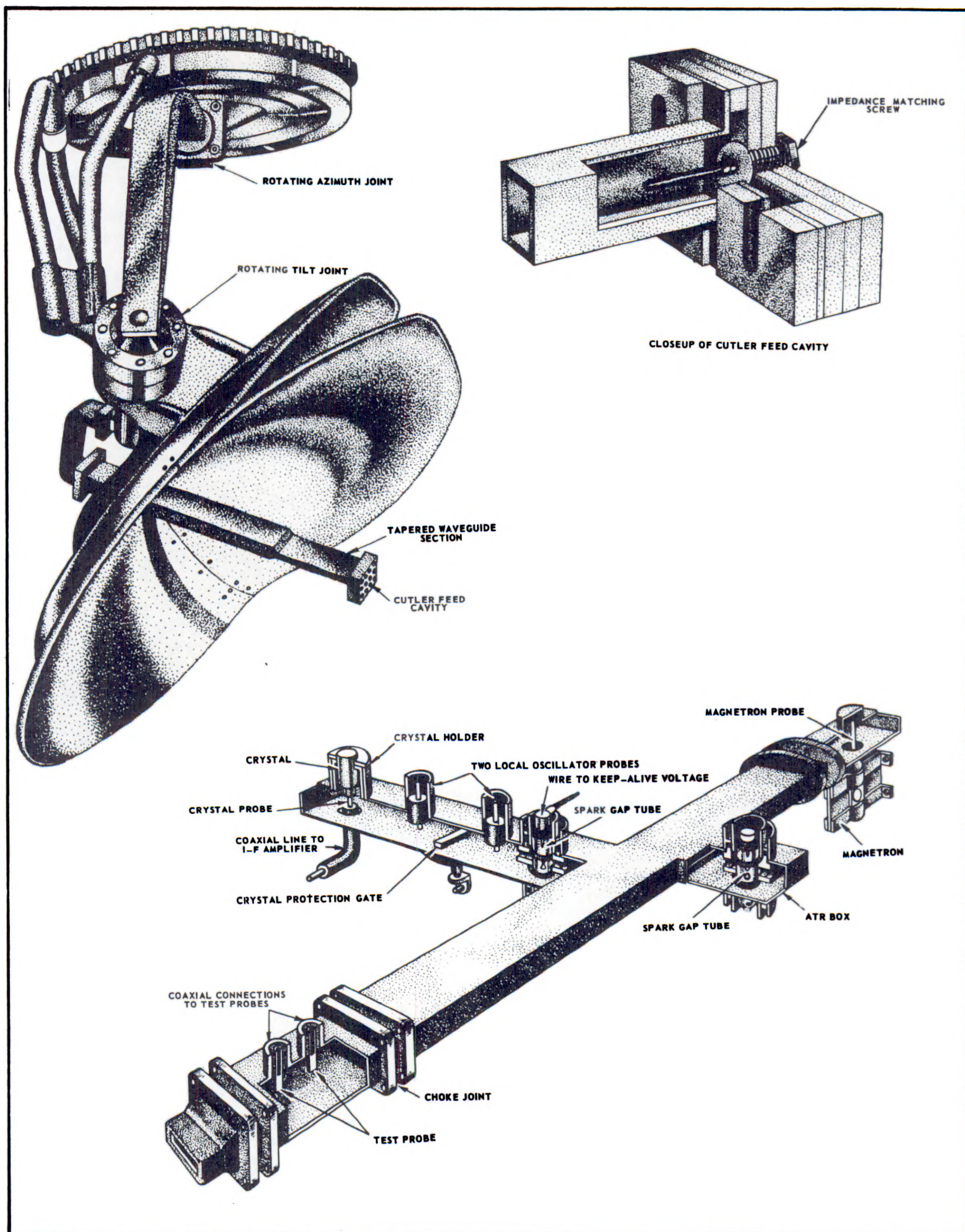


Figure 11-64. Complete Waveguide System

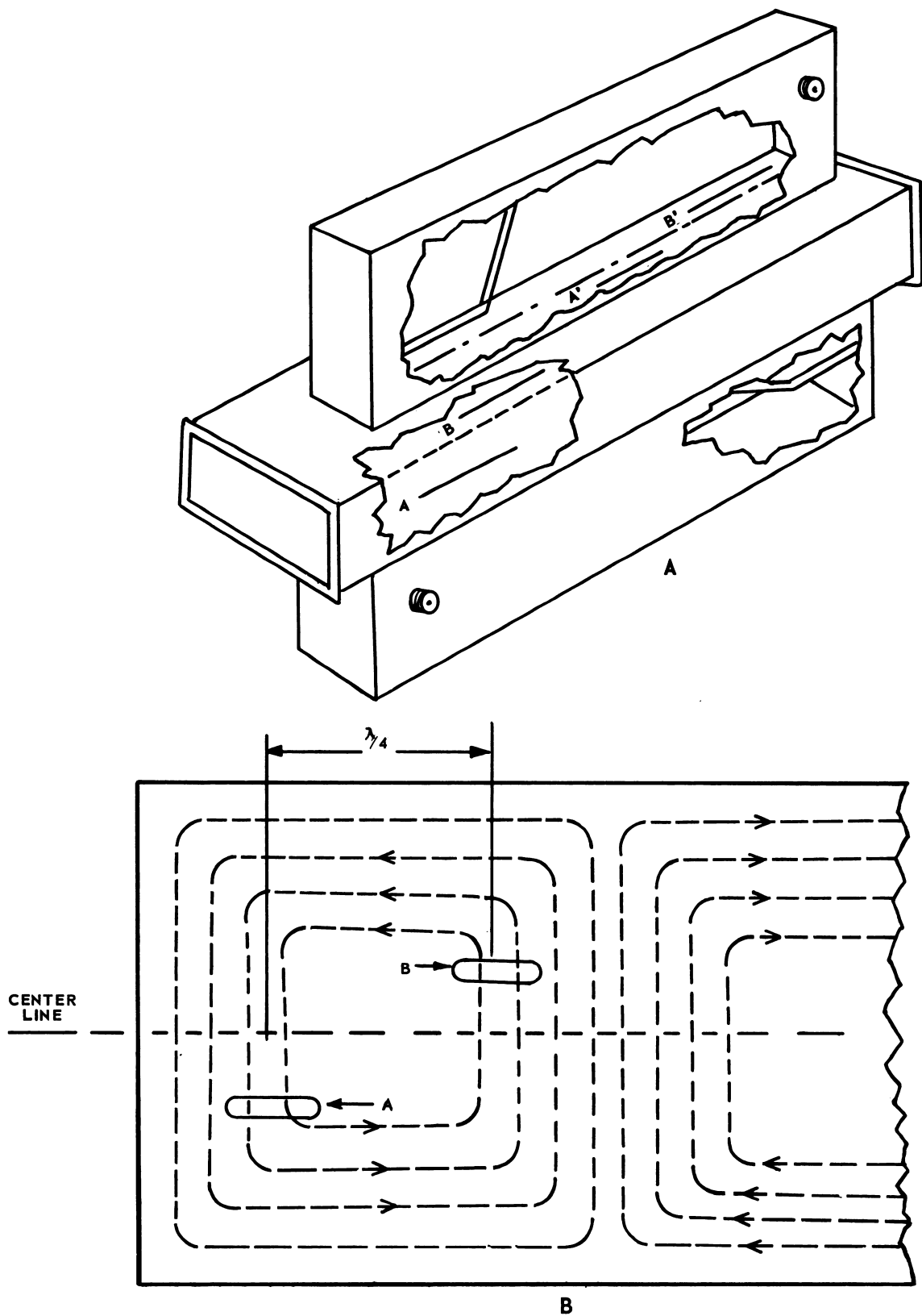


Figure 11-65. Bidirectional Coupler

spark gap is fired because the signal is too weak to strike an arc.

The signal enters the mixing chamber, where it strikes the end wall and reflections are again set up. The chamber becomes a cavity resonator. An additional signal is introduced from another oscillator — the heterodyne oscillator for the receiver. The signal is injected with a probe at the center of the E-field in the cavity. Both signals cause current flow through a pickup probe at a quarter-wave from the end of the cavity. This probe is connected to a crystal mixer. The second oscillator is set at a different frequency and produces the correct intermediate frequency when the returning signal is from a radar beacon transmitter.

As the crystal is easily damaged by high-powered signals, the TR box spark gap protection must be certain. To insure that it will arc over at once when the transmitter signal appears in the mixing chamber or when other nearby radar transmitters accidentally send in a weaker — but still damaging — signal, a *keep-alive* voltage is provided. This is a high DC voltage that causes the spark gap to be partially ionized. Then the additional voltage of any signal that is stronger than a radar echo will cause complete ionization or an arc between the electrodes. When the set is turned off, no keep-alive voltage is present, and the crystal is vulnerable to signals from other nearby radar transmitters. A spring-operated gate closes the waveguide when the set is turned off. No signal can get to the crystal from the antenna. When the set is turned on, an electrically operated relay opens the gate.

The entire system is assembled with rubber gaskets at the joints, or otherwise sealed to permit pressurizing the interior of the waveguide. The small windows or slots in the antenna cavity are covered with mica. It lets r-f energy through but keeps moisture out.

### DIRECTIONAL COUPLER

To measure the power from a radar transmitter, to measure the standing-wave ratio of a waveguide, or to feed a signal to a radar receiver, it is necessary to tap into the waveguide. The device used for this purpose is a directional coupler.

A typical coupler is shown in Figure 11-65A. This is a bidirectional coupler of the reverse type.

It consists of two couplers mounted on opposite voltage walls of the waveguide. The lower coupler taps off incident power; the upper taps off reflected power. Each coupler consists of an auxiliary waveguide section that receives energy from the main waveguide through two coupling slots. At one end of the auxiliary guide is a probe to couple energy to test equipment. At the other end is a load to dissipate unwanted power.

Note that the coupling slots are located on opposite sides of the center line of the waveguide. In addition, slot A is a quarter-wavelength farther along the guide than slot B. The magnetic field on one side of the center line of the waveguide is  $180^\circ$  out of phase with that on the opposite side. This is illustrated by Figure 15-65B. Since the slots are on opposite sides of the center line of the waveguide, they interrupt currents traveling in opposite directions, as shown by the arrows. For this reason, energy coupled into slot A is  $180^\circ$  out of phase with energy coupled into slot B.

Figure 11-66A illustrates the functioning of the coupler when it is used to tap off incident power.  $A_0$  is a point of reference, and is a quarter-wavelength from point B. The two parallel lines *do not* indicate parallel paths of power; rather, they are used to indicate energy in opposite halves of the waveguide. Keep in mind that the distance AD is the same as distance BC ( $\lambda/4$ ); their relative lengths appear different on the illustration because it is a two-dimensional drawing. A small amount of the energy traveling down the transmission line toward the antenna couples through slot A and travels a quarter-wavelength to point D. One quarter-wavelength farther down the line, energy couples through slot B. This energy travels path BCD to reach point D.

Energy reaching point D from slot A travels a quarter-wavelength (equivalent to  $90^\circ$ ). Energy reaching point D from slot B travels the distance from  $A_0$  to B ( $\lambda/4$ ), plus path BCD ( $\lambda/2$ ), the sum of which is equivalent to  $270^\circ$ . This results in a phase difference of  $180^\circ$  between the two signals. Add to this the  $180^\circ$  phase difference due to the energies being coupled from opposite sides of the center line of the waveguide, and the total phase difference is  $360^\circ$ . Since this represents a full cycle, the fields at point D are in phase, and power is available at the probe. Energy from slot A can also travel the path ADC

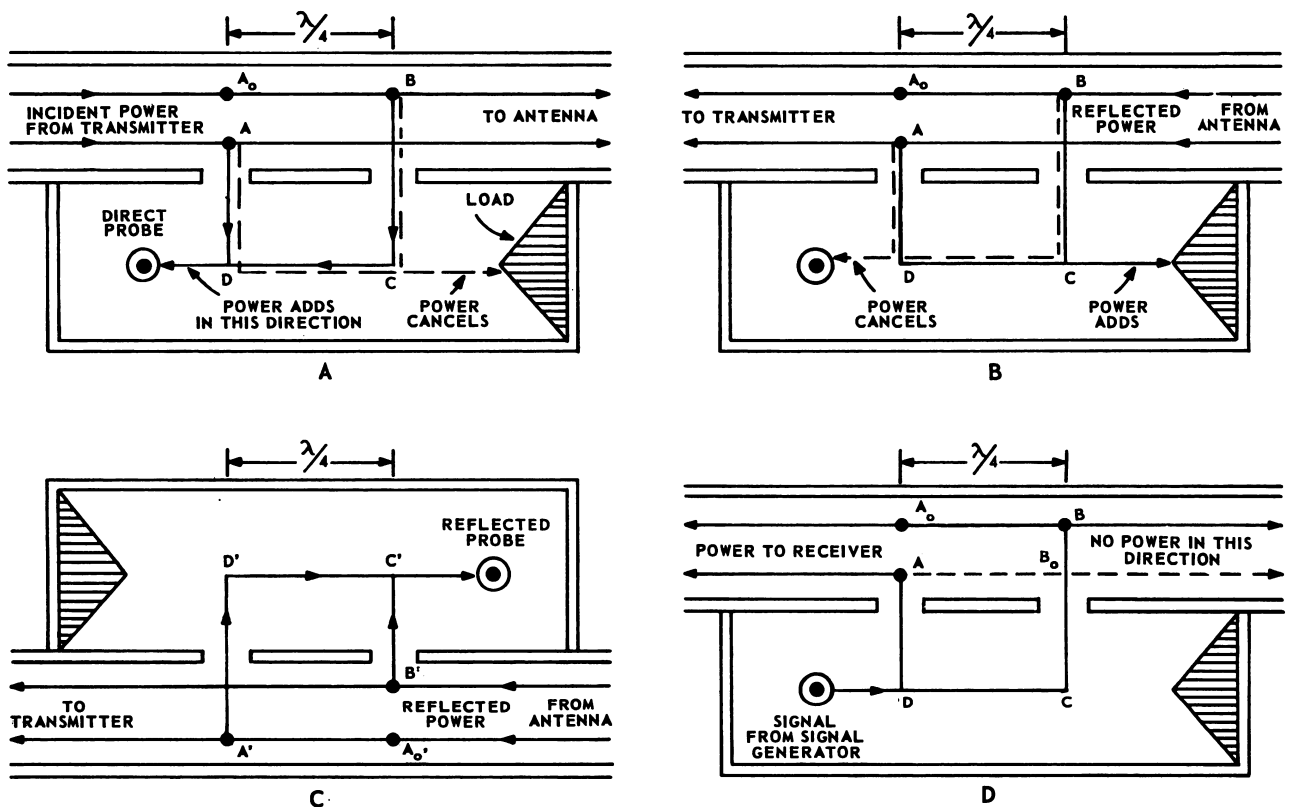


Figure 11-66. Principle of Operation of Bidirectional Coupler

( $\lambda/2$ ), and power from slot B can travel the path  $A_0BC$  ( $\lambda/2$ ). Since the paths are equal, and since the energy coupled through slot B is  $180^\circ$  out of phase with that from A, the fields cancel at point C.

Figure 11-66B illustrates the action of the coupler when reflected power is traveling along the waveguide in the opposite direction. In this case, the power components add toward the load in the coupler and are dissipated. Power cancels toward the probe; hence, no reflected power is measured. This illustrates the meaning of *directional coupler*, a device that is designed to measure power flowing in only one direction in the waveguide. With this coupler, therefore, it is possible to measure the incident power without its being affected by the standing-wave ratio. If the waveguide has a low standing-wave ratio, the incident power is a good measure of the power delivered by the transmitter.

In order to measure standing-wave ratio, it is necessary to measure both incident and reflected power. To accomplish this, a second coupler

must be used. It is placed on the opposite side of the waveguide and turned in the opposite direction. Note that this coupler is identical to the one just discussed, with the exception that the probe is in the opposite end of the coupler. (See Figure 11-66C.) The power coupled through slot  $B'$  is  $180^\circ$  out of phase with the power coupled through slot  $A'$ . However, power entering at slot  $A'$  must travel a half-wavelength farther, so it is shifted  $180^\circ$  by the time it reaches point  $C'$ . Therefore, the two fields arrive at  $C'$  in phase and add. As a result, reflected power is coupled to the probe.

Incident power will also enter these slots, but will cancel toward the probe and have no effect.

Figure 11-66D illustrates the feeding of a signal to the main waveguide through the lower coupler. The signal entering the coupler at the probe travels the paths shown to coupling slots A and B. At point A the energy coupled through slot A has traveled a quarter-wavelength ( $AD$ ). The energy is coupled into one side of the waveguide. By the time the energy coupled through



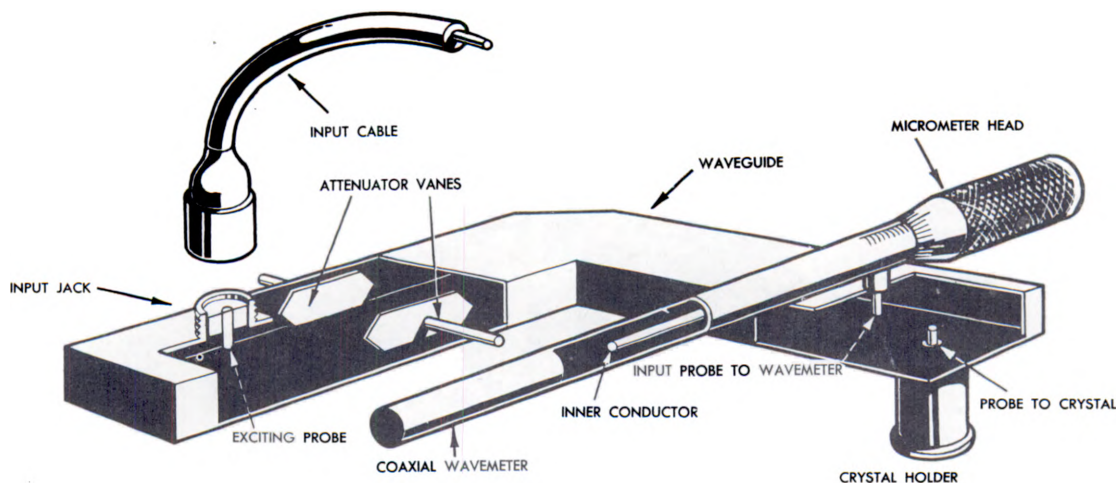


Figure 11-67. Accurate 3-cm Frequency Meter using a Waveguide

slot B reaches point  $A_0$ , it has traveled three-quarters of a wavelength ( $DCBA_0$ ). This is a half-wavelength farther than the energy coupled through slot A. It is, therefore,  $180^\circ$  out of phase with the energy coupled through slot A. But this energy is coupled into the opposite side of the waveguide, and hence it is in the correct phase relationship to excite the guide in its dominant mode. Therefore, the power will travel down the guide to the receiver.

The signal coupled through slots A and B will also attempt to travel toward the antenna to be lost as radiation. But no power can travel in this direction. Signal paths  $DAB_0$  and  $DCB$  are equal, hence power coupled into one side of the center line of the waveguide (through slot A) is in phase with that coupled into the opposite side through slot B. Thus, the waveguide cannot be excited in the dominant mode in this direction and no energy is lost through radiation.

### TEST EQUIPMENT

A resonant cavity can be used to measure the frequency of the signal in a waveguide. When approximate measurements are to be made, either the ring box or echo box can be used. When more accurate measurement is desired, a frequency meter of the type shown in Figure 11-67 is used. In this meter a coaxial cable brings the signal from a test jack in the waveguide of the radar set to the input jack of the wavemeter waveguide. A probe sends the signal down the

waveguide through attenuator vanes. When these vanes are at the walls of the waveguide they have no effect, but when they are moved together, they reduce the wide dimension until it is less than the cutoff wavelength for the frequency involved. The signal bounces back and forth between the vanes, with very little of it getting out of the waveguide proper. It then goes to a coaxial cavity.

This cavity is similar to that previously illustrated in Figure 11-53C, except that it has a micrometer head to insure maximum accuracy of reading. The coaxial line is resonant when it is 15 quarter-waves long in this test set. Resonance is indicated by the rectified output from the crystal, which is used to swing a meter; or if pulsed signal is used, the amplitude of the pulse can be observed on a test oscilloscope.

You can measure the standing waves on any waveguide system by inserting a slotted section into the waveguide as shown in Figure 11-68, and then using a travel-probe to measure the electric field strength. A special assembly fits on the slotted section onto the waveguide. This assembly picks up energy by a probe and transmits it through a short waveguide to a crystal, where it is rectified and then carried by a coaxial cable to an indicator, which is similar to that described earlier in the slotted coaxial line.

### ADJUSTMENTS

Although there are many adjustments in a waveguide system in any radar set, most of

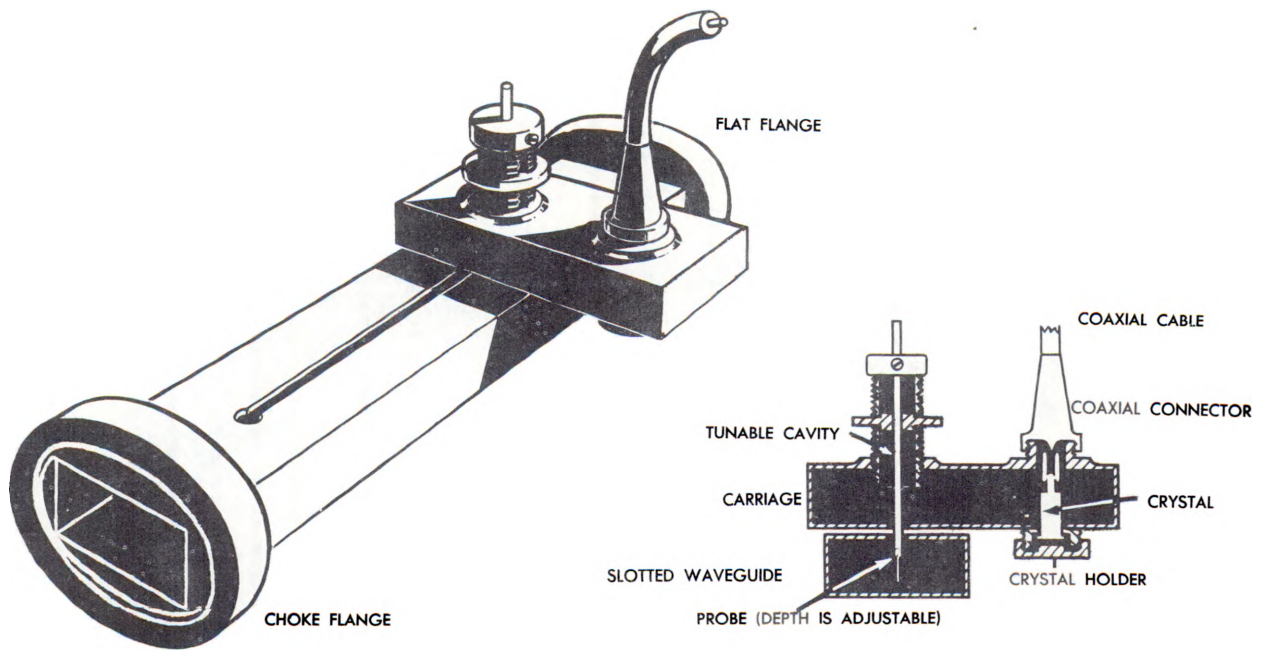


Figure 11-68. Slotted Waveguide Section Permits Access to Fields in Waveguide System

them are made at the factory. For example, almost all of the probes which a waveguide system uses cause an impedance mismatch. This mismatch must be tuned out. The plugs of the type shown in Figure 11-69 are provided to get rid of the mismatch, but adjustment is made at the factory where possible and the adjustment is fixed by soldering.

On the other hand, an adjustment which the radar mechanic does make deals with the amount of signal introduced by the local oscillator. This can be varied by making the probe in the mixing chamber longer or shorter. This probe, which is permanently fixed to the oscillator, moves the entire oscillator up and down with respect to the waveguide.

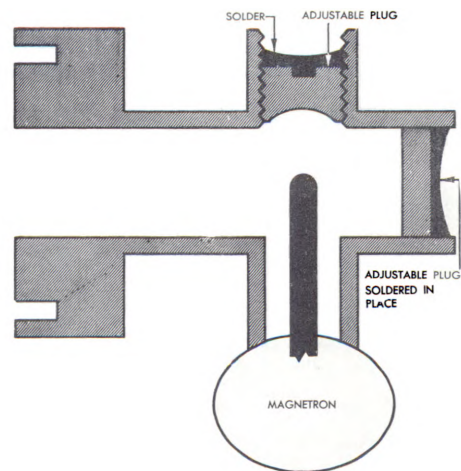
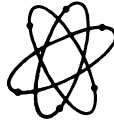


Figure 11-69. Mismatch and Reflections Introduced by Magnetron Probe are Tuned Out by Plugs





# **UHF Oscillators**

A radar transmitter consists of an oscillator and its associated pulsing circuits. The theory of pulsing circuits has been discussed to some extent in previous chapters. One purpose of this chapter is to acquaint you with oscillators and oscillatory circuits employed in radar transmitters; another is to give you additional information about pulsing circuits in transmitters and to acquaint you with local oscillators used in radar receivers. Specifically, it discusses problems involved in generating high-frequency signals, explains the theory and operation of a number of oscillators employed in radar equipment, and describes various methods for pulsing oscillators.

## **ADVANTAGES OF HIGHER FREQUENCIES**

Radar oscillators are operated at higher frequencies for three reasons: higher frequencies produce the best echoes, they make it possible to detect smaller targets, and they permit the use of small antennas.

As mentioned previously, radar transmitters send out short pulses at regular intervals. Usually, the duration of these pulses is a microsecond or less to prevent the echo from any particular pulse from being returned by nearby targets before that pulse ends at the transmitter. Furthermore, it is proved by experiments that satisfactory echoes require that the transmitted pulse consist of 200 or more oscillations. Obviously, then, the only way for a narrow pulse to contain 200 oscillations is to operate the transmitter oscillator at ultra-high or microwave frequencies; for ex-

ample, 200 million oscillations per second for a 1-microsecond pulse.

The size of the target that radar can detect depends mostly on the frequency. In general the higher the frequency, the smaller the target that the radar set can detect. A target gives best echoes when it is about a half-wave, or some multiple, of the frequency being transmitted. If the target is much smaller than a half-wave, the intensity of pulse that it reflects is low. However, as the operating frequency is increased, the half-wave becomes smaller and makes it possible to detect smaller targets.

Perhaps the main reason for using higher frequencies is that it is possible to construct directive antennas in a smaller space than when lower frequencies are used. Small directive antennas are an especially important consideration in airborne equipment where space and weight are major problems.

## **FREQUENCY LIMITATIONS OF USUAL OSCILLATORS**

The principal factors which limit the frequency of the ordinary type oscillator are the construction of the tube, and the external circuits that connect to the tube.

### **Tube Construction**

The tube construction factors that limit the operating frequency of the usual, or conventional oscillator are the interelectrode capacitances in

the tube, the inductances of the leads, and the transit time.

**INTERELECTRODE CAPACITANCES.** At ordinary radio frequencies the interelectrode capacitances in a vacuum tube have reactances which are so large that they do not cause any serious trouble. However, as frequencies increase, the reactance of these capacitances become small enough to materially affect the performance of a circuit. A 1-mmf capacitor, for example, has a reactance of 159,000 ohms at 1 mc. If this capacitor is the plate-to-grid capacitance and the r-f voltage between these electrodes is 500 volts, there will be an interelectrode capacitance current flow of  $500/159000$  or .00315 ampere (3.15 ma), an amount of current that will not seriously disturb circuit operation. On the other hand, at 100 mc the reactance of this capacitor becomes 1590 ohms and the current flow is  $500/1590$  or 0.315 ampere, an amount which will seriously affect circuit performance.

A good point to remember is that the higher the frequency or the larger the interelectrode capacitance, the higher the current flow through this capacitance. In most UHF oscillators, interelectrode capacitance currents are much greater than the power currents supplied by the tube. These higher currents cause power losses in the resistance in the oscillatory circuit.

Since interelectrode capacitances are effectively in parallel with the tuned circuit, they affect the frequency at which the tuned circuit resonates. As you can see in Figure 12-1, the plate-to-cathode capacitance is in parallel with the series combination of the plate-to-grid capacitance and the grid-to-cathode capacitance. All these capacitances together form a part of the total capacity of the tuned circuit. When the tube is operating, the effective capacitance varies due to the Miller effect.

Interelectrode capacitance limits the frequency by establishing a minimum capacitance below which it is impossible to go. In addition, interelectrode capacitance varies with the applied voltages and with the loading of the oscillator. This causes frequency instability, particularly when the interelectrode capacitance forms a large part of the tuning capacitance.

**INDUCTANCE OF LEADS.** Another frequency-limiting factor within a tube is the inductance of the leads to each tube element. While these

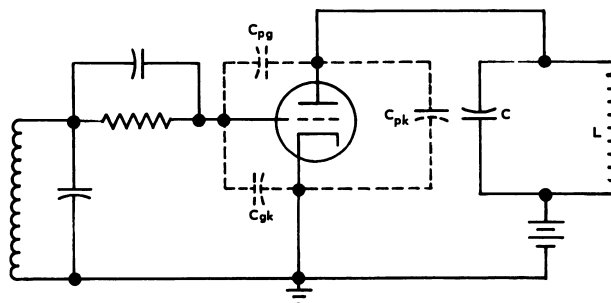


Figure 12-1. Effect of Interelectrode Capacitance on Frequency of TPTG Oscillator

inductances do not necessarily impair the efficiency of the oscillator, they may represent a major portion of the inductance of the tuned circuit and limit the frequency by setting a minimum limit on the inductance. Furthermore, since the cathode lead is common to both plate and grid circuits, feedback takes place through it and produces an additional loss of efficiency.

**TRANSIT TIME.** A third limitation imposed by tube construction is *transit time*. Transit time is the time required for electrons to travel from cathode to plate. At lower frequencies transit time is negligible, since it occupies only a comparatively small portion of an oscillatory period. But as the frequency becomes higher, transit time occupies an appreciable portion of this period and produces undesirable effects in tube operation. The effect of transit time is of special concern in connection with the input impedance of the tube.

Part of the current that flows in the grid circuit is the current which charges the grid-to-plate capacitance,  $C_{gp}$ . The voltage that produces this current is the vector sum of the input voltage (grid-to-cathode), and the output voltage across the plate load. At lower frequencies, with a resistive load, these two voltages are  $180^\circ$  out of phase and add algebraically to determine the charging current. This current is  $90^\circ$  out of phase with the input voltage. But at higher frequencies where transit time is an important factor, the plate current begins to lag the input voltage. This causes the plate voltage to be less than  $180^\circ$  out of phase with the input voltage, and the voltage across the capacitor to lag the input voltage slightly. As a result, the charging current will no longer be  $90^\circ$  out of phase but

will have an in-phase component. This means that power is consumed in the grid circuit. This consumption of power is effectively the same as adding a high resistance in the input impedance. This resistance decreases as the frequency increases.

**OVERCOMING TUBE LIMITATIONS.** There are several ways to reduce the effects of interelectrode capacitances in vacuum tubes. None are completely satisfactory. One is to move the electrodes farther apart. However, this is hardly desirable because it increases the transit time. Another method is to reduce the size of the tube and the electrodes. This is satisfactory except for one ill effect. It decreases the power handling ability of the tube directly with the square of the factor by which the electrodes are reduced. Another method is to separate the leads and to bring them out of the envelope at the nearest point. This results in a slight decrease in the capacitances.

Similarly, there are several ways to reduce the inductances of the leads. As just mentioned, bringing out electrode leads through the envelope at the nearest point produces a slight decrease in the electrode capacitance in a tube. This also decreases the inductances of the leads. Another method is to make double connections to the electrode. This makes two parallel inductances which cuts the lead inductance in half. Another method is to arrange the leads as extensions of external transmission lines.

For reducing transit time, closer spacing of electrodes and higher plate voltages are used. Closer spacing of electrodes, however, produces higher interelectrode capacitances. Therefore, this method can be used only when the electrodes are made smaller. For this reason radar oscillators operate at very high plate voltages. The fact that a radar transmitter does not operate continuously is an advantage in that high enough voltages may be used to reduce transit time without exceeding the maximum power rating of the tube. Operating with high voltage does, however, necessitate some precautionary measures, such as separating the seals of the leads to avoid the presence of excessive voltage gradients in or on the surface of the glass envelope, and avoiding sharp projections on the high-voltage leads or electrodes to prevent arcing.

## Limitations Due to External Circuits

Increases in frequency cause a quite rapid increase in the power losses in the external oscillatory circuit. These power losses are due principally to skin effect and radiation.

**SKIN EFFECT.** Skin effect causes a considerable increase in the resistance in a vacuum tube circuit. This results in a lower  $Q$  and increased  $I^2R$  losses. To prevent skin effect losses, conductors are made large in size and tubular in shape since current flows only in the surface. In addition, they are plated with silver, since it has a higher conductivity than copper, the material out of which most conductors are made.

**RADIATION.** Radiation, the other cause of power loss, is due to incomplete cancellation of electromagnetic fields in the region surrounding the circuit. When the frequency is low enough so that the spacing between two parallel conductors equals only a very small fraction of a half-wavelength, there is almost complete cancellation of fields in all directions. At higher frequencies, however, an identical spacing would represent a larger fraction of a half-wavelength. This means less cancellation.

In extreme cases, such as where the spacing is a half-wavelength, the fields add in the direction of the plane containing the two conductors. This causes the tuned circuit to radiate energy as does an antenna. Therefore, as the frequency increases, it is necessary to reduce the spacing between the parallel elements. But there is a limit on how far you can reduce the spacing. Too close spacing, for example, causes arcing, which materially increases the r-f resistance of the tank circuit. Another means of eliminating radiation is using concentric lines instead of open wire lines. This eliminates radiation entirely, since the outer conductor acts as a shield, preventing the electromagnetic field from expanding into space.

**LIMIT OF INDUCTANCE AND CAPACITANCE.** At UHF frequencies it is necessary that the inductances and the capacitances in the oscillatory circuit be very small. The limit of capacitance is the sum of the interelectrode capacitances and the distributed capacitance of the leads. The limit of inductance is the lead inductance plus the inductance necessary to connect the tube electrodes externally. UHF oscillators approach both these limits, the only capacitance in the

tank circuit being a small trimmer capacitor for fine tuning adjustments, and the inductance being a short-circuited transmission line less than a quarter-wavelength in length.

Figure 12-2 illustrates a typical parallel line tank circuit. The expression for the input reactance (looking at the circuit from the tube) of the tank circuit is

$$Z = jR_c \tan B_s$$

where  $R_c$  is the characteristic resistance of the line and  $B_s$  the electrical length.

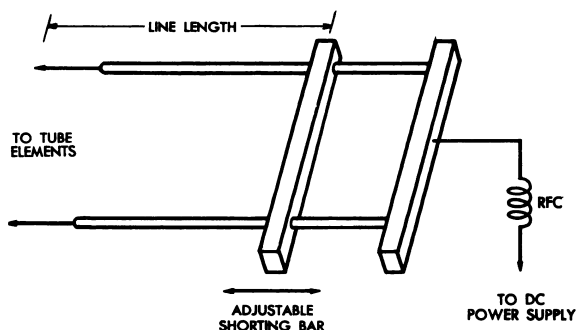


Figure 12-2. Typical Parallel-Wire Resonant Circuit

As you can see in the expression, if  $B_s$  is less than  $90^\circ$  ( $\lambda/4$ ), then  $\tan B_s$  is positive and the reactance is inductive. For smaller values of inductance the length of the line is decreased. Further, this expression is true only when the resistance losses are low enough to be neglected. Frequently the tank circuit is located near a flat surface and parallel to it. The flat surface acts as a shield and, while it does not affect the electrical length of the line, it does lower the characteristic impedance of the line.

### TRIODE OSCILLATORS AS TRANSMITTERS

A variety of oscillators are used in radio and radar transmitter equipment. While this discussion deals primarily with their use in the transmitting circuits of this equipment, it is also applicable to local oscillators in receivers. The main difference between the two kinds of oscillators is that the radar transmitter oscillator is pulsed and produces considerable power during each pulse, while the local oscillator operates continuously at a lower power output than the transmitter oscillator.

### Ultraudion Oscillator

The typical ultraudion oscillator in Figure 12-3 is used as a local oscillator in some radar receivers, and as a transmitter in certain radar sets. The circuit shown is incapable of generating enough power to be useful for high-power long-range sets.

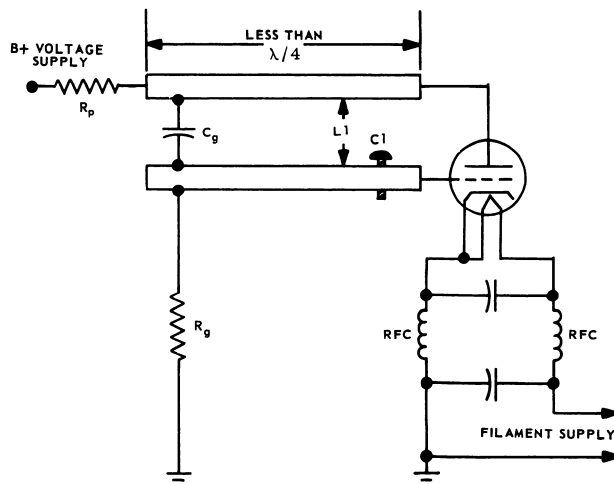


Figure 12-3. Ultraudion Oscillator

The DC circuits consist of the components,  $R_p$ ,  $C_g$ ,  $R_g$ , and the small DC resistances in the tank circuit and leads. The components,  $C_g$  and  $R_g$ , form the grid-leak bias circuit. The charge built up on  $C_g$  biases the tube to class C operation.

To understand the operation of the ultraudion oscillator, consider its equivalent circuits shown in Figure 12-4. So far as r-f is concerned, in the equivalent circuit the chokes in the cathode are open circuits. The tank circuit inductance  $L$  consists of the plate and grid lead inductances connected in series with the parallel combination of  $L_1$  and  $C_1$ .

The simplified equivalent shows that this tank circuit inductance forms a resonant circuit with the combined interelectrode capacitance,  $C$ , which is given by the equation,

$$C = C_{sp} + \frac{C_{gk} \times C_{pk}}{C_{gk} + C_{pk}}$$

The expression for the frequency of this resonant circuit is,

$$f_o = \frac{1}{2\pi\sqrt{LC}}$$

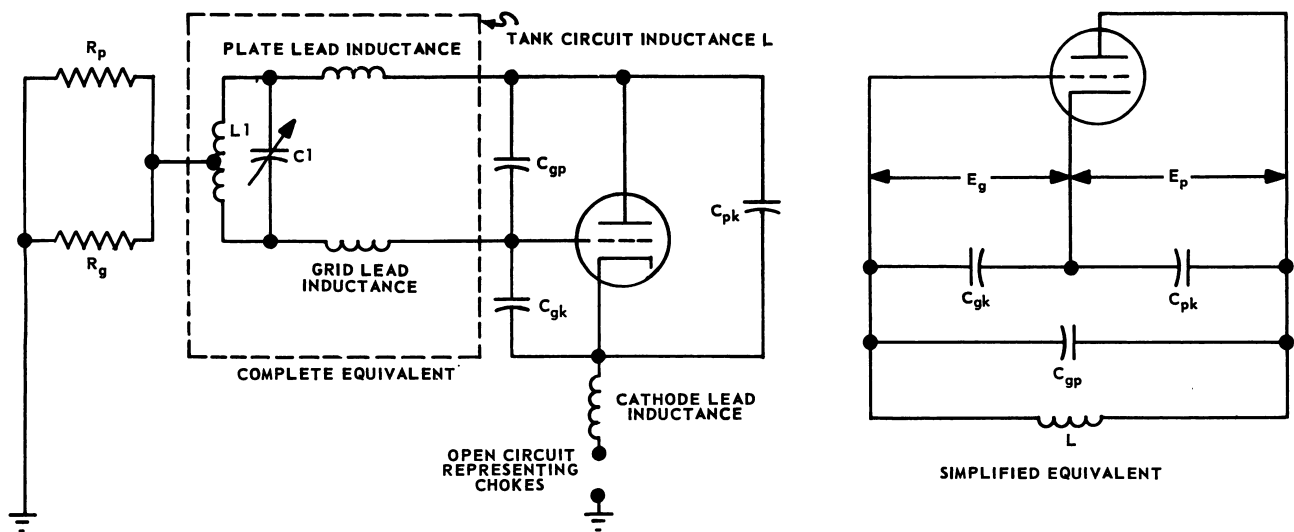


Figure 12-4. Equivalent Circuits of Ultraudion Oscillator for R-F Currents

Basically, the circuit of the ultraudion oscillator (see the simplified equivalent) is that of a Colpitts oscillator. The capacitances  $C_{gk}$  and  $C_{pk}$  form a voltage divider circuit. The voltage  $E_g$  developed across  $C_{gk}$  is in the right phase and high enough to sustain oscillations. The minimum amplification required to sustain oscillations is given by the formula  $A = C_{gk}/C_{pk}$ . If the initial gain is greater than this value, oscillations build up to such an amplitude that the gain is reduced to the value given by this formula.

### Lighthouse Tube Oscillator

The lighthouse tube oscillator like the ultraudion oscillator is a single tube oscillator. It is used in a few low-powered radar sets such as short-range gun-laying equipment and in some types of transportable beacon equipment, and also as a local oscillator in ground radar equipment. In a typical application the lighthouse tube, which gets its name from its peculiar construction, operates at a frequency of 2500 mc.

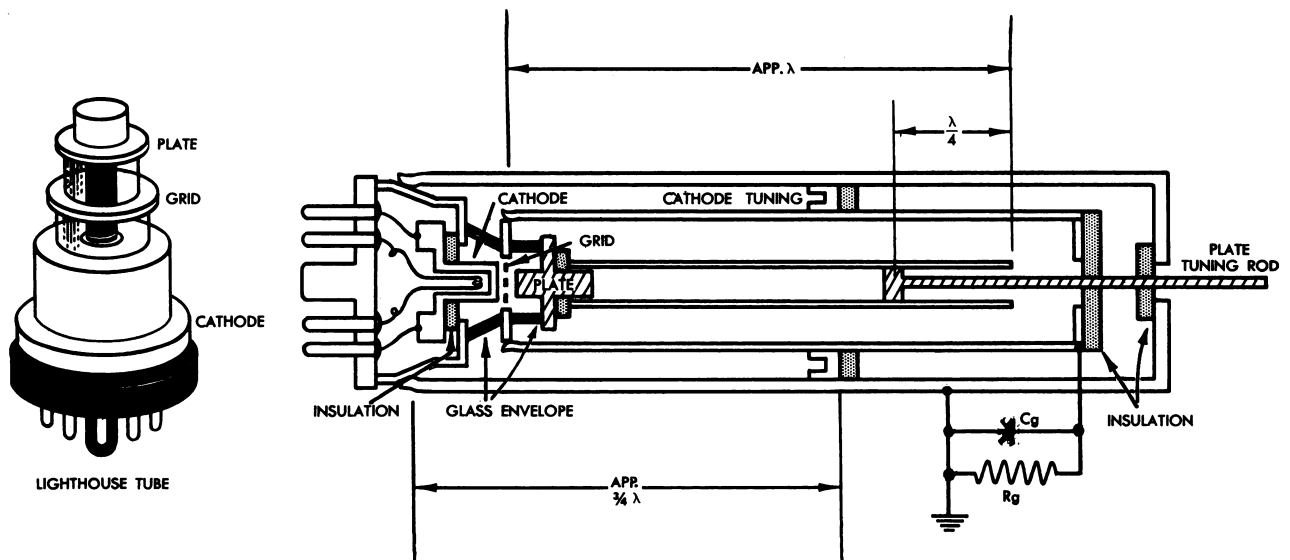


Figure 12-5. Lighthouse Tube Oscillator

Its plate, grid and cathode connections are arranged so that it may be mounted directly in the tuning assembly, as illustrated in Figure 12-5.

The component called the tuner consists of three coaxial cylindrical conductors. The inner cylinder makes contact with the plate, the next with the grid, and the outer conductor with the shell of the tube. The outer conductor provides an r-f connection to the cathode through capacitance. The space between the cathode and the grid conductors forms a coaxial cathode line which is shorted by a plunger. This plunger does not touch the grid conductor since this would form a DC connection from grid to cathode. The capacitance between the plunger and the grid conductor is great enough to form a virtual short circuit for r-f. The DC grid-to-cathode path, consisting of  $R_g$  and  $C_g$ , forms the grid-leak bias circuit.

In the plate circuit, the plate line consists of the grid and plate conductors. It has an open circuit at the end away from the tube. The plate voltage is applied to the inside of the plate line through the plate tuning rod at a point which is a quarter-wavelength from the end. This quarter-wave section, which is shorted at the point where B+ is applied, presents a high impedance to r-f from the open-ended plate line and, therefore, prevents r-f from flowing into the power supply.

**EQUIVALENT CIRCUITS.** The following equivalent circuits, Figure 12-6, show the evolution of the lighthouse tube oscillator. In the first equivalent circuit the components,  $L_k$  and  $L_p$ , represent the inductances of the shorted cathode line and the open-circuited plate line respectively. This is a true representation of the actual circuit,

provided that the cathode line is less than three-quarters of a wavelength long and provided that the plate line is less than a full wavelength long.

The three-quarter wavelength and full wavelength were chosen for this circuit rather than the one-quarter and one-half wavelength because at the frequency of operation the smaller fractions of a wavelength are inconveniently short. However, keep in mind that if the cathode line is slightly less than a quarter-wavelength, it appears as an inductance and that if the plate line is slightly less than a half-wavelength, it appears as an inductance. There is no lead inductance shown in the circuit since the lines connect directly to the electrodes. The interelectrode capacitances complete the equivalent circuit.

In the equivalent circuit, Figure 12-6B, the parallel combination of  $L_p$  and  $C_{gp}$  are represented by  $X_2$  and the parallel combination of  $L_k$  and  $C_{gk}$  by  $X_1$ . Obviously the reactances  $X_2$  and  $X_1$  along with  $C_{pk}$ , which you can consider unknown reactances, must form a resonant circuit. The reactance  $X_1$  forms a voltage divider circuit with  $C_{pk}$ . Since the voltage across  $X_1$  is the voltage fed back to the grid, it must be  $180^\circ$  out of phase with the plate voltage, which can happen only if  $X_1$  is a capacitive reactance. This condition leads to the third equivalent circuit, Figure 12-6C, in which the reactance  $X_1$  is represented by a capacitor.

An oscillatory circuit requires at least one inductance; therefore,  $X_2$  must be an inductive reactance. Hence in the third equivalent circuit (C) the reactance  $X_2$  is shown as the inductance  $L$ . This circuit is representative of the oscillator circuit only at the resonant frequency. Examine this circuit and note that basically it is a Colpitts

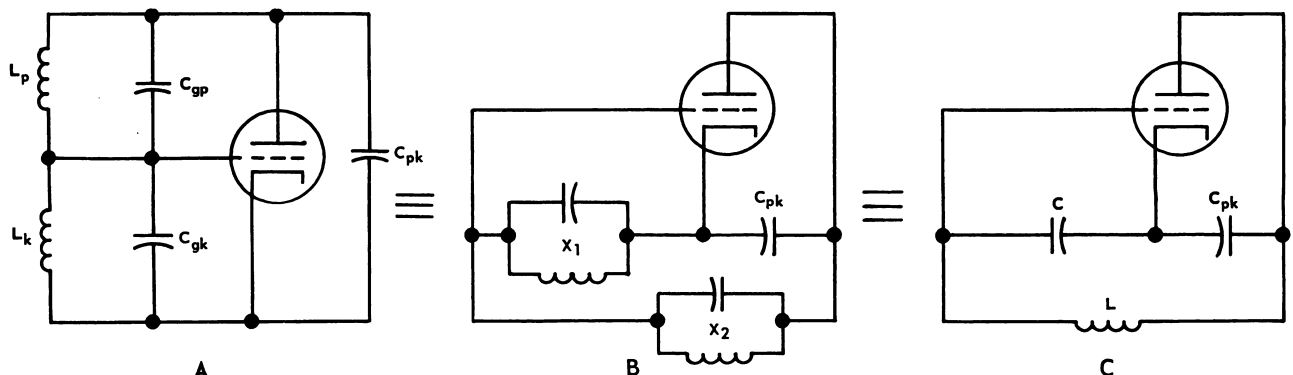


Figure 12-6. Evolution of Equivalent Circuit of Lighthouse Tube Oscillator



oscillator. Now, if  $X_1$  is capacitive,  $C_{gk}$  must conduct more heavily than  $L_k$ ; therefore the oscillator frequency must be above the resonant frequency of  $L_k$  and  $C_{gk}$ . Likewise, for  $X_2$  to be inductive, the oscillator frequency must be below the resonant frequency of  $L_p$  and  $C_{gp}$ . Thus, you see that the oscillator frequency is between the resonant frequencies of the plate circuit and the cathode circuit.

**FEEDBACK.** In this oscillator the amount of feedback depends upon the size of  $C$  in relation to  $C_{pk}$ . This is, in turn, a function of the tuning of the cathode line; hence, the cathode tuning serves principally for controlling the amount of feedback. If feedback is too small, oscillations will be weak and may cease completely. If feedback is too great, power consumption in the grid circuit will become too large and will result in a decrease in output of the oscillator. The value of  $C$ , hence of the cathode tuning, has some effect on the frequency of oscillations, but since  $C$  and  $C_{pk}$  are in series and  $C$  is usually much larger than  $C_{pk}$ ,  $C$  must be changed considerably to affect the frequency appreciably.

**PLATE TUNING.** Since the tuning of the plate line determines the resonant frequency of the plate circuit which, in turn, determines the amount of inductance represented by the plate circuit at the oscillator frequency, the plate tuning is the chief control of the oscillator frequency. It has a secondary effect of feedback, since  $C$  is determined by the relative frequencies of the cathode circuit and of the oscillator. You can see from the foregoing discussion that the plate line is tuned to give the correct frequency, and the cathode line to adjust feedback for maximum output from the oscillator. Since there is some interaction, however, it will be necessary to readjust each line for maximum accuracy.

### Push-Pull Oscillators

To obtain higher power outputs, two tubes in push-pull circuits, or four or more tubes in a ring oscillator are sometimes used. In push-pull circuits, the tuned circuits are the parallel line (Lecher line) type. This arrangement forms a balanced-to-ground system.

The oscillator circuit discussed first is the push-pull tuned-plate tuned-grid oscillator. Only the oscillatory action is described here. Its pulsing circuits are taken up later in this chapter.

**TUNED-PLATE TUNED-GRID OSCILLATOR.** In the tuned-plate tuned-grid oscillator, Figure 12-7, the plate and the grid lines are slightly under  $\lambda/4$  in length. They are indicated in the circuit of the tuned-plate tuned-grid oscillator as inductances  $L_p$  and  $L_g$ . The length of the plate line in the circuit may be varied by the shorting bar. While the grid line is of fixed length in this particular arrangement, a little greater flexibility would result if it, too, were adjustable in length.

The inductances  $L$  and capacitors  $C_1$ ,  $C_2$ ,  $C_3$ , and  $C_4$  in each of the cathode circuits form filters to prevent r-f from reaching the filament transformer. The grid-leak bias circuit includes  $R_g$  and  $C_g$ . The component  $C_g$  is shown as a dotted portion since it is actually the distributed capacitance of the grid line to ground. B+ voltage is applied to the plate lines through  $R_s$  and  $L_s$ .  $L_s$  serves the double purpose of preventing r-f from getting into the power supply and of suppressing parasitic oscillations.

In the equivalent circuit of Figure 12-8A, both tubes of the oscillator are shown.  $L_p$  is the inductance representing both the plate line and the plate lead inductances.  $L_g$  represents the inductance of the grid line and grid leads.  $L_k$  is the cathode lead inductance. With proper adjustment of  $C_k$ ,  $L_k$  forms a series resonant circuit that effectively grounds the cathode within the tube.

The second equivalent circuit, Figure 12-8B, shows only one tube since both circuits are identical.  $L_{pk}$  is one-half  $L_p$  and with  $C_{pk}$  forms a circuit that resonates at a frequency above that at which the entire circuit oscillates.

In the equivalent circuit, Figure 12-8C, the parallel combination of  $L_{pk}$  and  $C_{pk}$  is represented as an inductance  $L_2$ . Likewise  $L_{gk}$ , which is one-half  $L_g$ , with  $C_{gk}$  forms a circuit that resonates at a higher frequency than that of the oscillator. This circuit is represented as an inductance at the oscillator frequency. This is true because the tank circuit must have at least one inductive element to oscillate and the two reactances must be alike for the feedback to be in the correct phase. In this equivalent circuit the oscillator is the Hartley type. Since the oscillator frequency must be lower than either resonant frequency (plate or grid circuits), the grid or plate circuit, whichever is lower in resonant frequency, controls the oscillator frequency and the other controls the amount of feedback.



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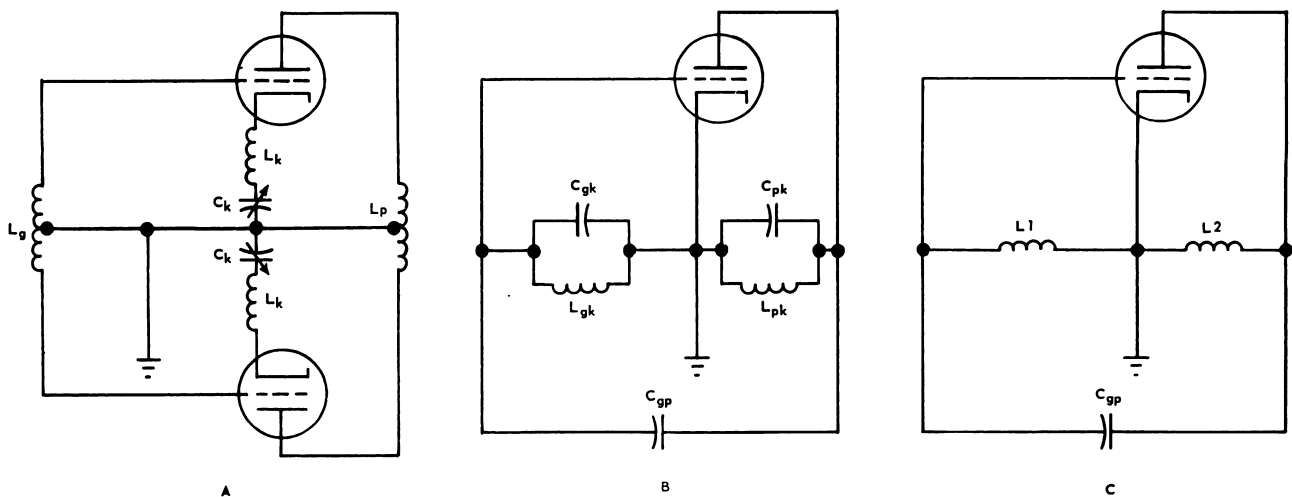


Figure 12-8. Equivalent Circuits of TPTG Oscillator

**TUNED-GRID TUNED-CATHODE OSCILLATOR.** Another push-pull oscillator circuit that is used in several types of UHF radar equipment is the tuned-grid tuned-cathode circuit, Figure 12-9. In this oscillator the plates are effectively short-circuited by tubes that have very little plate lead inductance and that are connected by a section of line. When the lead inductance is large enough to become objectionable, then the plates are connected by a half-wave section of line to insure a short circuit. The DC circuits are the same as in the tuned-plate tuned-grid oscillator.

In the first equivalent circuit, Figure 12-10A, both tubes are shown. The grid line is represented as inductance  $L_g$ , the cathode line as reactance  $X_k$ , and the interelectrode capacitances as indicated.

The circuit at B shows only one tube. In this circuit  $L_{gp}$  represents one half of  $L_g$  and  $X_{pk}$  represents one half of  $X_k$ . By following the line of reasoning similar to that used in the TPTG circuit, you see that the parallel combination of  $X_{pk}$  and  $C_{pk}$  must be capacitive at the oscillator frequency. Thus, it is shown as capacitor  $C_k$  in the equivalent circuit at C.

The oscillator must have an inductive element; hence  $L_{gp}$  and  $C_{gp}$  in parallel act as an inductance  $L$ . For  $L_{gp}$  and  $C_{gp}$  to be inductive, the grid line must always be less than a quarter-wavelength. Since variations in  $L$  have the greatest effect on the frequency of oscillations, *the grid line is the primary frequency tuning device*. The cathode line ( $X_{pk}$ ) controls only a portion of the total

capacitance of the circuit and therefore does not greatly affect the oscillator frequency. However, since  $X_{pk}$ , in parallel with  $C_{pk}$ , forms a voltage divider with  $C_{gk}$ , it has a *considerable effect on feedback*.  $X_{pk}$  may be inductive, purely resistive, or capacitive, depending on whether the cathode line length is less than, equal to, or greater than a quarter-wavelength at the frequency of oscillation.

If  $X_{pk}$  is a high resistance, the feedback is determined entirely by the voltage divider composed of  $C_{gk}$  and  $C_{pk}$ . If  $X_{pk}$  is capacitive (greater than a quarter-wavelength), then the two capacitances in parallel combine to increase  $C_k$ . This results in an increase in feedback. Tuning the cathode line to slightly less than a quarter-wavelength makes it inductive and decreases  $C_k$ , which results in a reduction of feedback. Oscillations will cease if  $X_{pk}$  and  $C_{pk}$  in parallel can no longer be represented as a capacitor. As can be seen by the third equivalent circuit, this oscillator is basically a Colpitts oscillator.

The tuned-grid tuned-cathode type oscillator has three advantages. First, the cathode line offers a convenient means of connecting the load to the oscillator because it is at ground potential with reference to DC. Second, since only DC voltages appear at the plate and only r-f voltages appear in the cathode circuit, the two are not present in the same circuit and the high peak voltages present in a tuned plate circuit are avoided. Third, the tuning of the oscillator is quite simple, the grid line being the frequency

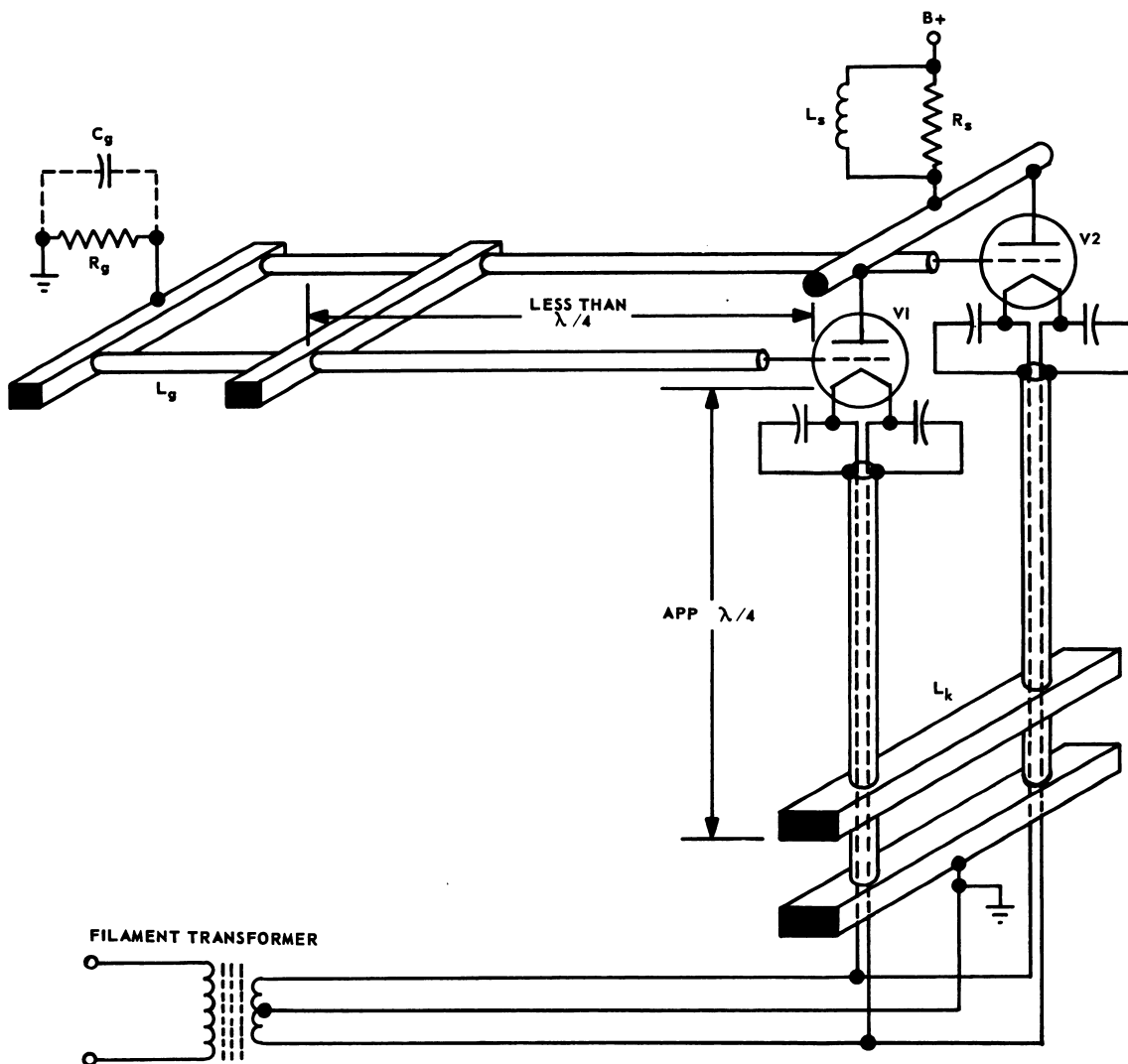


Figure 12-9. Tuned-Grid Tuned-Cathode Oscillator

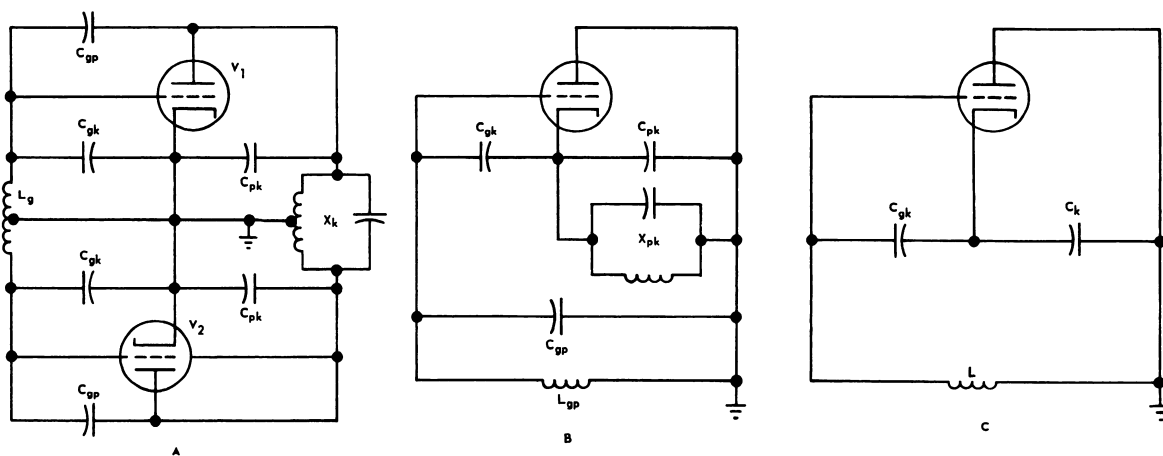


Figure 12-10. Equivalent Circuits of Tuned-Grid Tuned-Cathode Oscillator

control and the cathode line the feedback control. There is some interaction between the tuning controls so that readjustment of each may improve both accuracy and output.

### Tuned-Plate, Tuned-Grid, Tuned-Cathode Oscillator Circuit

A number of radar oscillators in the UHF range employ tuned lines in all three leads of

the triode, as in the oscillator circuit in Figure 12-11. So far as r-f is concerned, all three tuned lines are grounded at their midpoints. The DC circuits and filament circuits are identical to those described previously. In the circuit shown, the plate and grid lines, which are less than a quarter-wavelength, are labelled  $L_p$  and  $L_g$  respectively. The cathode line, which may be equal to, less than, or greater than a quarter-wavelength, is labelled  $X_k$ .

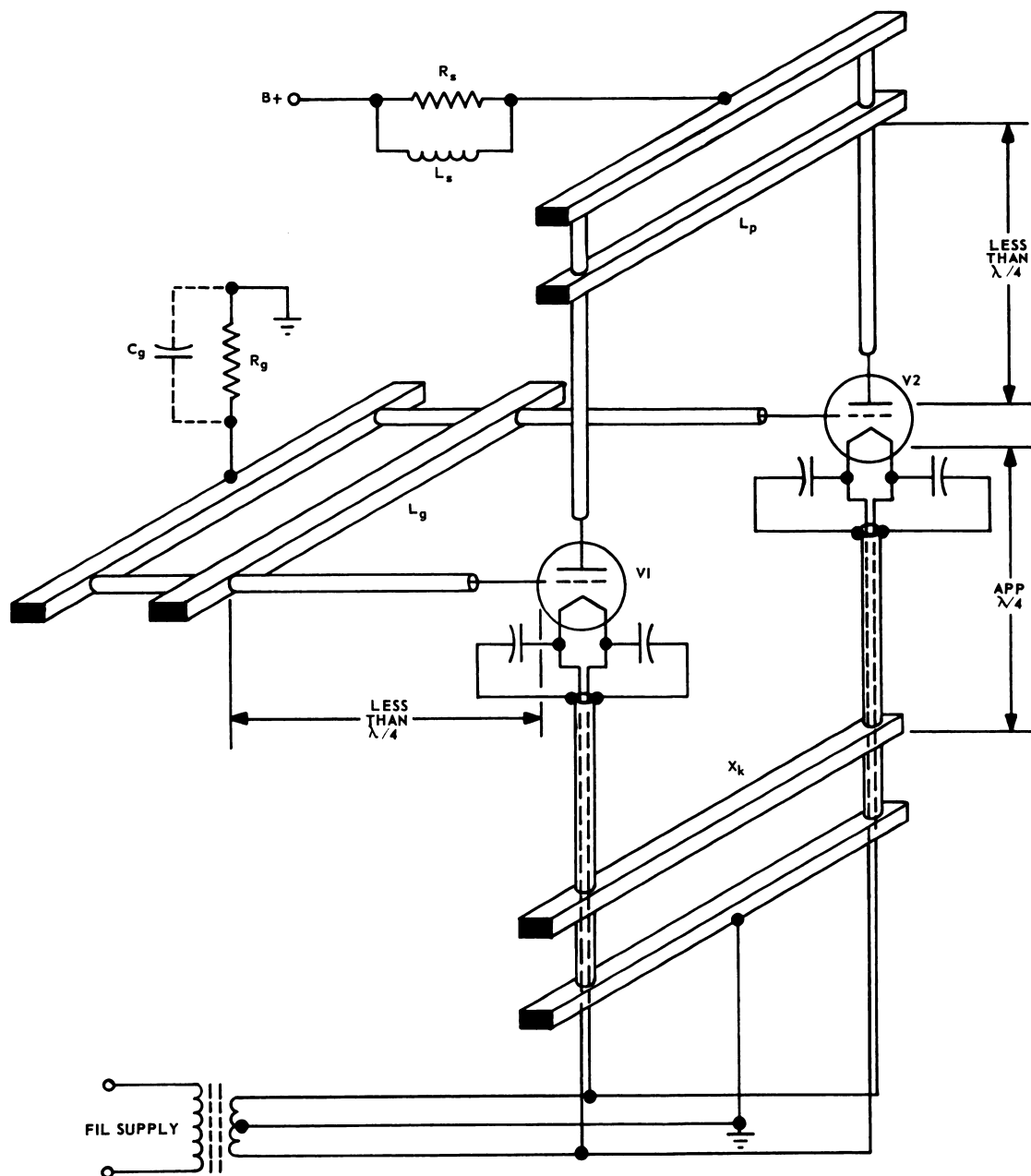


Figure 12-11. Tuned-Plate, Tuned-Grid, Tuned-Cathode Oscillator Circuit

Equivalent circuit, Figure 12-12A, shows both tubes, the inductances  $L_p$  and  $L_g$  center-tapped to ground, and the unknown reactance  $X_k$  center-tapped to ground. The interelectrode capacitances are not shown in A. Since the circuits of the two tubes are identical, the second equivalent circuit shows only one tube. The inductances  $L_{pn}$  and  $L_{gn}$  and the reactance  $X_{kn}$  represent one-half of  $L_p$ ,  $L_g$ , and  $X_k$ . The interelectrode capacitances are added.

Examination of the circuit at B shows that if the cathode line is exactly a quarter-wavelength so that  $X_{kn}$  becomes a high resistance, then the circuit is that of a conventional Colpitts oscillator. If the ratio of  $L_{gn}$  to  $L_{pn}$  is the same as the ratio of the reactances of  $C_{gk}$  and  $C_{pk}$ , there is no difference in potential between the cathode and ground, and  $X_{kn}$  is in effect out of the circuit. This condition also reduces the circuit to that of a Colpitts oscillator. In either case the frequency of oscillation is determined by the sum of the inductances of the plate and grid lines, and the amount of feedback by the ratio of  $C_{gk}$  and  $C_{pk}$ . Ordinarily this ratio does not give the amount of feedback necessary to cause oscillations for best efficiency; therefore, the cathode line is detuned from the pure resistive quarter-wavelength so that  $X_{kn}$  is either capacitive or inductive.

Making  $X_{kn}$  capacitive or inductive while keeping the sum of  $L_{gn}$  and  $L_{pn}$  constant does not

change the frequency of oscillation but does increase feedback. The exact amount of feedback depends on the ratio between  $L_{gn}$  and  $L_{pn}$ . For example, consider the case where  $L_{pn}$  is reduced to zero and examine the nature of  $X_{kn}$ . Here  $X_{kn}$  is in parallel with  $C_{pk}$  as shown in the equivalent circuit of Figure 12-10B. Thus, in controlling feedback, the effect of  $X_{kn}$  changes is the same as in the tuned-grid tuned-cathode oscillator—that is, the cathode line must be increased in length to increase feedback or be decreased in length to decrease feedback.

On the other hand, if  $L_{gn}$  is decreased to zero and  $L_{pn}$  is increased to equal the value of their former sum, then  $X_{kn}$  is in parallel with  $C_{gk}$ . Its effect on feedback then is just the opposite to that described. For example, if the cathode line is greater in length than a quarter-wave, it becomes capacitive, and the parallel combination  $C_{gk}$  and  $X_{kn}$  becomes a larger capacitance. This reduces the feedback. Reducing the length of the cathode line below a quarter-wavelength makes it inductive, and the parallel combination becomes a smaller capacitance, resulting in an increase in feedback.

This demonstrates that *feedback is a function of both the ratio of  $L_{gn}$  and  $L_{pn}$  and of the tuning of the cathode line*. Since the frequency of oscillation is controlled by the sum of  $L_{pn}$  and  $L_{gn}$  and feedback is controlled both by their ratio

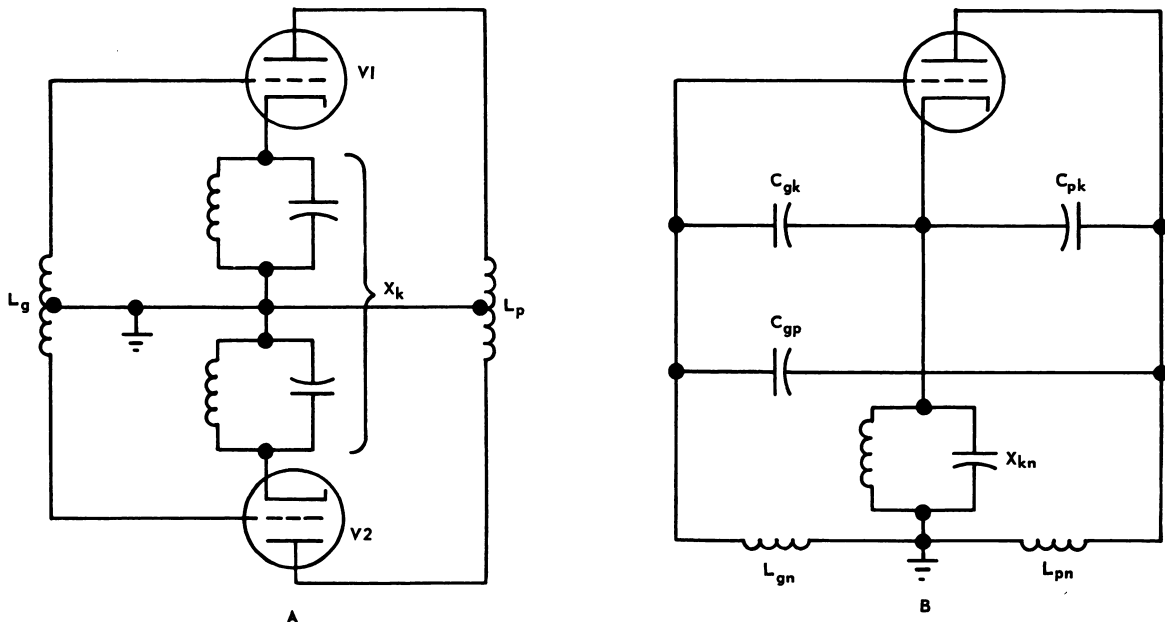


Figure 12-12. Equivalent Circuits of TPTGTK Oscillator



and by the length of the cathode line, there are numerous settings of the three variable line lengths for any given frequency of oscillation and for proper feedback. This is readily understood if you consider that you have three variable elements with which to control two variables, and that there is a certain amount of interaction among the three controls.

### Pulsing of Triode Oscillators

There are two methods of pulsing UHF oscillators. One method uses separate modulator tubes; the other employs a self-pulsing oscillator.

**THE SELF-PULSING OSCILLATOR.** Most oscillators use grid-leak bias or a combination of grid-leak and fixed bias. An advantage of grid-leak bias is that the amplitude of oscillations is limited and stabilized. Although grid-leak bias was taken up in Chapter 4, it is discussed briefly again to introduce the subject of self-pulsing. In this connection notice the graphs of plate current and grid voltage versus time in Figure 12-13. These waveshapes are the same for all oscillators that use grid-leak bias. At zero time plate voltage is applied, but the grids are grounded so that a steady plate current flows. At time  $t_1$  the ground is removed, and any random change in plate current will cause a change in voltage in the plate tank circuit.

A portion of this change is fed back to the grid circuit. Since this feedback is in the right phase and amplitude, the oscillations will grow in amplitude. Now each time the grid goes positive, grid current flows and charges the grid-leak capacitor more negatively. On each successive oscillation the average grid voltage (the bias) becomes more negative. This continues until a state of balance is reached where just enough grid current is drawn to replace the charge that leaks off through the grid-leak resistor during the remainder of the cycle. In addition, as the average grid voltage becomes more negative, the plate current flows for a shorter portion of each cycle until the flow is just sufficient to replace the energy lost in the tank circuit and to supply feedback to the grid circuit.

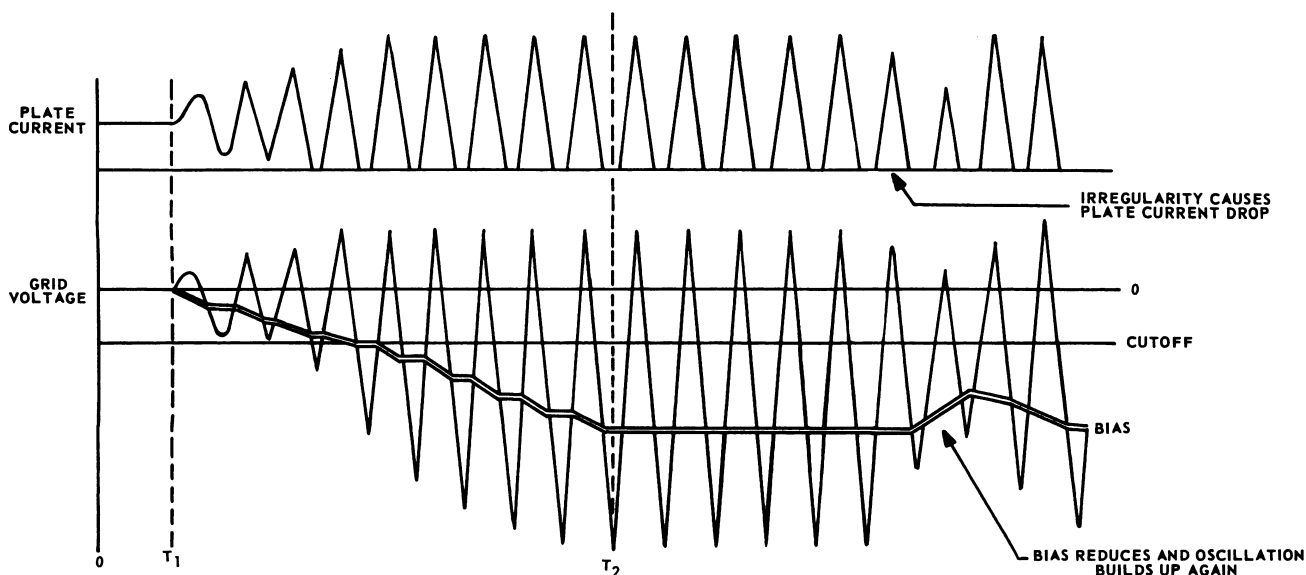
With perfect circuit elements and smooth electron emission, the circuit would oscillate indefinitely at this amplitude. But minute irregularities occur in the movement of electrons through resistors and conductors; in addition, emission

from a cathode is always in nonuniform bunches. The nonuniformity is slight, but enough to affect the equilibrium in the oscillator circuit. Any slight decrease in current decreases the feedback, and if the time constant of the RC circuit is not too long, the bias is reduced slightly by discharge of the capacitor. The decrease in bias allows a little more plate current, which re-establishes the equilibrium at the old level.

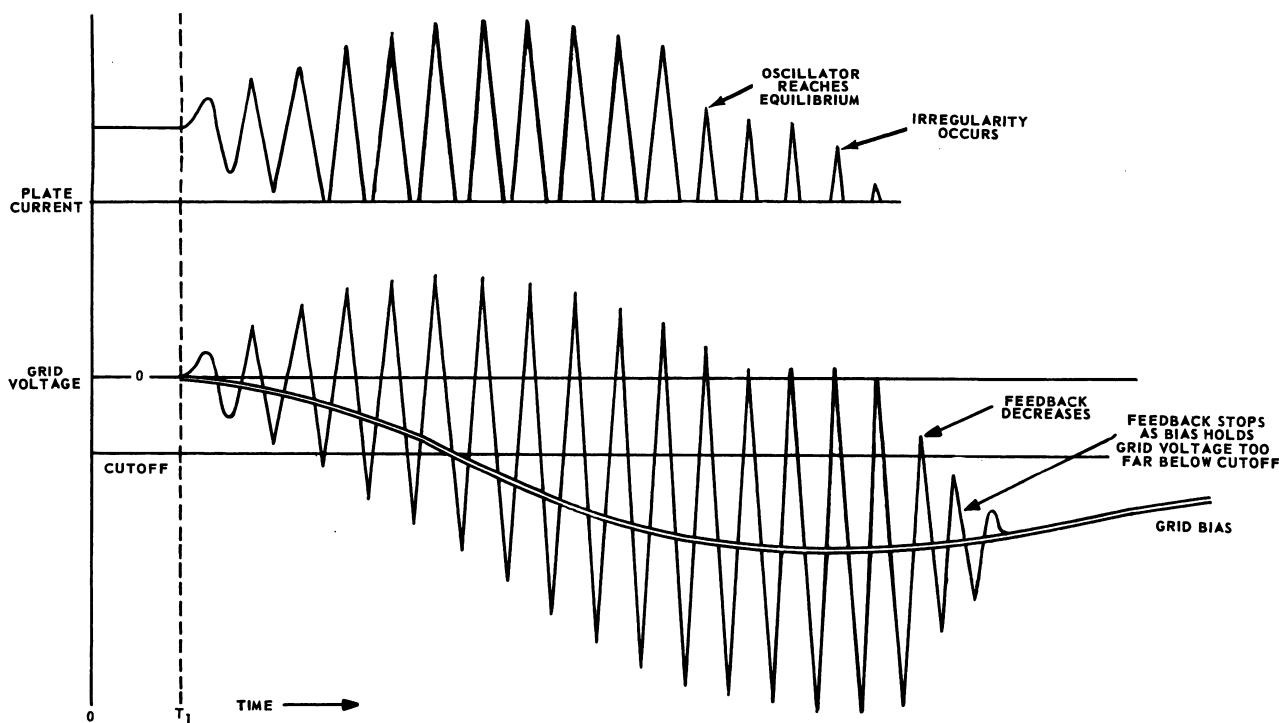
In a circuit with a long RC time constant, however, the circuit starts oscillating, then settles down to a balanced condition as previously described. But any slight decrease in plate current now will not be compensated for by a lowered bias, as shown in Figure 12-13B; the capacitor discharges too slowly. The resulting continued high bias prevents the feedback from being restored to the proper level and after a few cycles, the oscillation dies out completely. The capacitor slowly discharges, and when its voltage reaches cutoff, current again flows and the process repeats.

For continuous operation, then, the RC time constant of the self-biasing circuit must be short enough to allow automatic adjustment of the bias for changing conditions in the circuit. To make the circuit oscillate briefly, then stop, use a long RC time constant. You can see that the time the tube remains cut off depends upon both the grid capacitor and the grid resistance. By properly choosing these values, you can have the oscillator operate for one or two microseconds and then cease operation for 500 to 1000 microseconds, as you desire. You can get an idea of how the variations of grid voltage would look from Figure 12-14A. It is not possible to show the actual time relations since the "off" period is several hundred times as long as the "on" period.

In a radar set that uses a self-pulsing oscillator, synchronizing pulses are taken from the oscillator for starting the sweep on the CRT and for any other circuits that need to be synchronized. In most cases the exact pulse recurrence frequency is unimportant. There are a few sets, however, which use the 400-cps power supply to regulate the PRF. This is done by applying a small synchronizing voltage to the grid as shown in Figure 12-14B and C. Without synchronization the grid voltage rises exponentially and crosses the cutoff line at a small angle. This



A PLATE CURRENT AND GRID VOLTAGE VS TIME SHOWING BUILD UP OF OSCILLATIONS



B EFFECT OF LONG RC

Figure 12-13. Plate Current and Grid Voltage in Oscillators

means that the actual time that oscillations begin may vary. Notice that in C the point at which the grid voltage crosses the cutoff bias line is much more definite. Other ways of controlling the PRF more exactly are to feed in a positive-going pulse when the grid voltage approaches

cutoff or to return the grid to a positive voltage rather than to ground.

In order to make the duration of the pulse more definite, some self-pulsed oscillators use an artificial transmission line instead of the grid-leak capacitor. This has the added advantage of

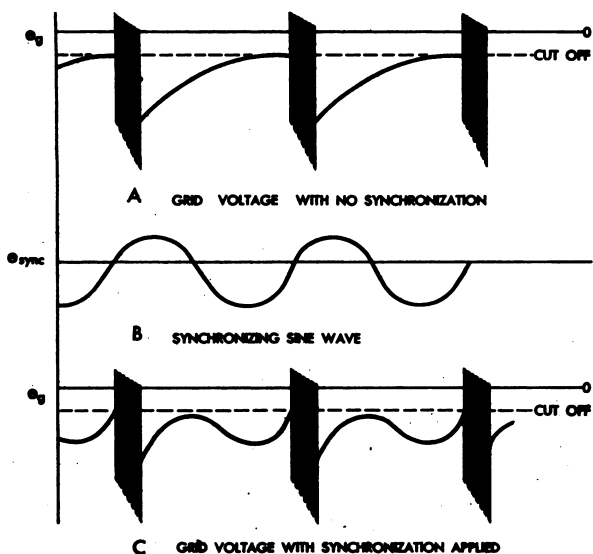


Figure 12-14. Grid Waveshape With and Without Synchronization

making the operating conditions more nearly constant throughout the pulse. There are two places where the pulse line can be located. One is to put it in the grid circuit as shown in Figure 12-15; the other is in the cathode circuit. (The construction and operation of artificial transmission lines was explained in Chapter 10.)

A pulse line behaves as a long transmission line. Your chief interest in it is that it acts as a resistive circuit until the charging pulse has

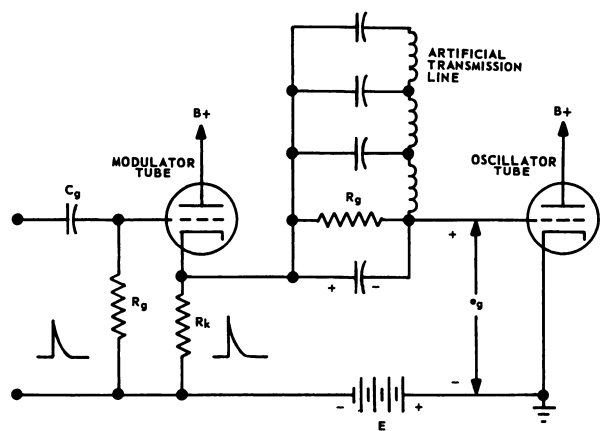


Figure 12-15. Oscillator Circuit with Pulse Line in Grid Circuit

traveled its length and returned, at which time the line is fully charged. This causes the voltage to change in two steps, the first from the bias voltage to the voltage determined by the product of the grid current times the characteristic impedance of the line. The second step is of equal magnitude. The PRF is controlled by the synchronizing pulses, but the discharge of the pulse line through the grid resistor must be such that the grid voltage approaches the cutoff point by the time the next synchronizing pulse comes in.

The discharge of the line is in steps, as you can see in Figure 12-16. The magnitude of the

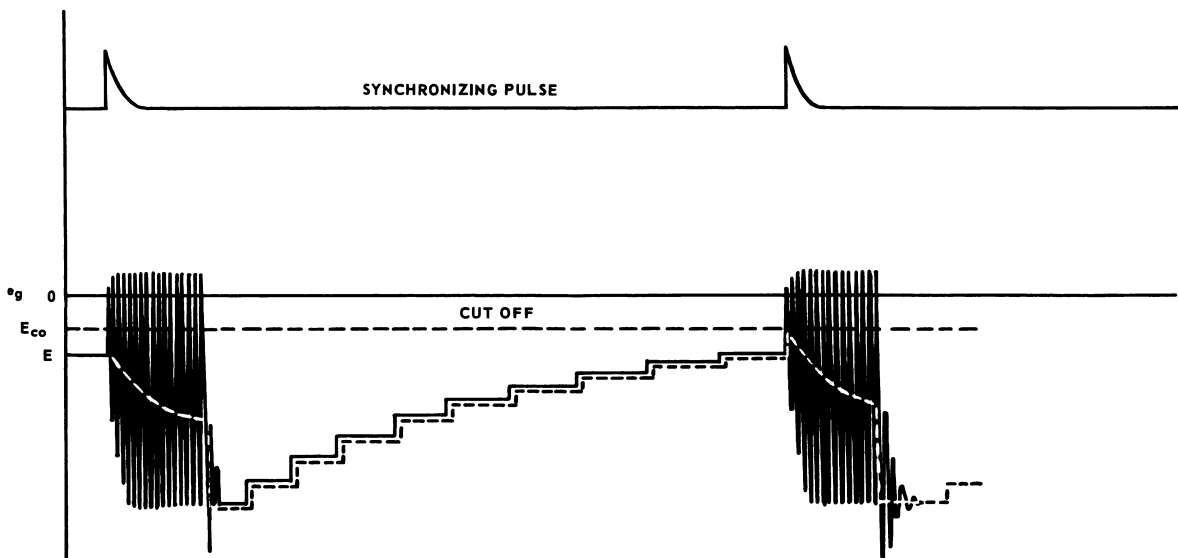


Figure 12-16. Grid Waveshape of a Line-Controlled Self-Pulsed Oscillator

discharge voltage is determined by the relative values of  $R_g$  and the characteristic impedance of the line.  $R_g$  is usually much larger, hence the steps are small. These steps follow the general path of an exponential discharge curve with a time constant of  $R_g C_o$ , where  $C_o$  is the capacitance of the pulse line.

**SEPARATE MODULATORS.** As mentioned earlier, some triode UHF oscillators are not self-pulsed, but use separate modulators. These modulators are very much like those used with magnetron oscillators which are discussed later in the chapter. Here we merely note that the oscillator may be plate-modulated, grid-modulated, or cathode-modulated. In Figure 12-17, short, positive, rectangular pulses are applied to the grid of the modulator tube resulting in a negative-going pulse at the plate. The transformer is connected so that plate voltage is applied to the oscillator tube, or tubes, in short rectangular pulses. R-f components for the oscillator are not shown in the diagram.

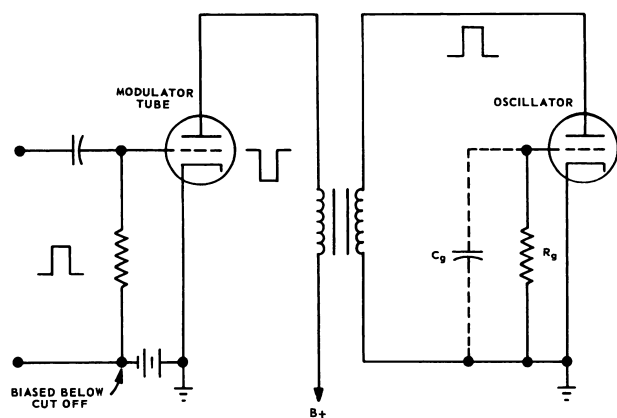


Figure 12-17. Plate-Modulated Oscillator

The transformer is specially constructed so that it passes the rectangular pulse without undue distortion. Since the purpose of the transformer is polarity inversion, why not apply negative-going pulses to the grid of the modulator tube to cut it off and get a positive-going pulse at the plate for use in modulating the oscillator? This would require that the modulator tube, or tubes, conduct heavily during all the cycle except during the pulse and would result in excessive waste of power.

The grid capacitance  $C_g$  of the oscillator is quite small, usually consisting of the capacitance-to-ground of the grid-tuned circuit. If made too large,  $C_g$  might cause the oscillator to be self-pulsed and cut itself off before the modulator pulse was over. Hence  $C_g$  and  $R_g$  have values to cause the oscillator to operate continuously if plate voltage were applied all the time. The modulator pulse then is most effective in controlling the output pulse. Oscillations must build up rapidly so that the leading edge of the output pulse will be as steep as possible. The plate voltage of the oscillator must remain constant throughout the pulse.

A cathode-pulsed (modulated) oscillator, Figure 12-18, does away with the necessity of inverting the pulse, hence no transformer is used. The charging diode serves to recharge the coupling capacitor between pulses. The RC time constant of the capacitor discharge path must be quite large so that the capacitor will not discharge appreciably during the pulse and thereby lower the plate-to-cathode voltage.

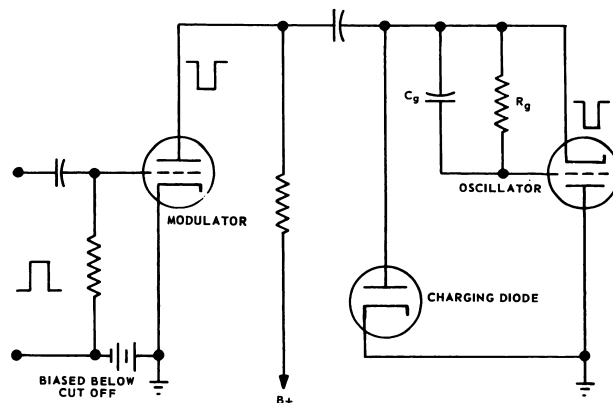


Figure 12-18. Cathode-Modulated Oscillator

Grid modulation requires less power from the modulator for operation than either of the other two types of modulation. It uses a pulse which has not been amplified. In grid modulation, Figure 12-19, a positive-going rectangular pulse is applied to the grid which raises the grid voltage well above the cutoff point, and oscillations start. Oscillations stop when the grid voltage drops to the normal bias level at the end of the pulse.  $C_g$  in the grid-pulse oscillator shown is a small capacitance which is in addition to the grid-to-ground

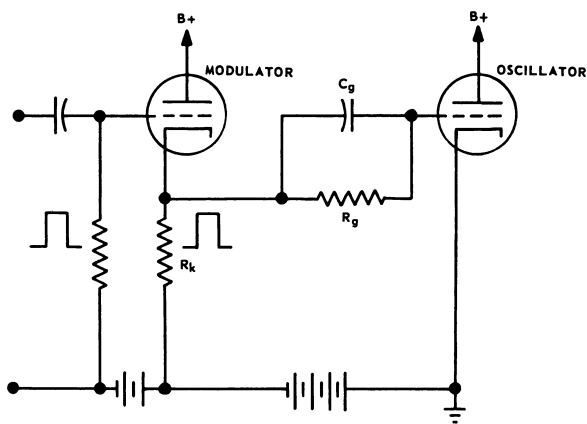


Figure 12-19. Grid-Modulated Oscillator

capacitance. Its purpose is to couple the pulse from the cathode of the modulator tube to the grid of the oscillator to insure a rapid rise of voltage at the grid and cause oscillations to build up quickly.

### THE MAGNETRON OSCILLATOR AS A TRANSMITTER

Before the days of modern radar, electronic equipment operated at considerably lower frequencies. One reason was because a transmitter could not be constructed to operate in the UHF band and produce a usable output signal. These low frequency transmitters made a large, bulky antenna necessary and it wasn't possible to design an accurate lightweight radar set. Then, with the invention of the magnetron, came a new era in radar. Here was a lightweight, high-frequency oscillator that could produce a usable output with a power level that was unheard of before. The magnetron can produce output signals of up to 30,000 megacycles and power levels up to 6 megawatts.

This discovery was what the military electronics field was looking for. Small, lightweight radar sets with very narrow beams could now be built. Of course, with this discovery came new problems. New circuits had to be designed to operate the magnetrons. New types of modulators came into existence. The line-pulsing modulator seemed to be made for the magnetron. New filament supply circuits were designed.

The main advantage of the magnetron as an oscillator tube is that transit time, which is a

determining factor in the upper limit of frequency of operation in conventional tubes, is not a disturbing influence. There are some special types of triodes used in unusual transit time oscillator circuits that will oscillate in the frequency range useful for radar, but unlike the magnetron their power output is too small for them to be useful in most cases. In addition, there are also the velocity-modulated tubes, such as the Klystron and the Shepherd Pierce tube, which operate in the desired frequency range with better power output than the triodes just mentioned.

The magnetron is a diode in which the magnetic field between the cathode and the plate is perpendicular to the electric field. The tuned circuits included in the tube are in the form of cylindrical cavities. As shown in the typical 10-cm magnetron in Figure 12-20, the cathode and filament structure is at the center of the tube. It is supported by filament leads which are large and stiff enough to keep the cathode and filament structure fixed in position under ordinary circumstances.

Exercise care in handling the magnetron tube, since undue jarring of the tube or bending of the filament leads might change the position of the cathode causing it to short to the plate, or change its spacing with the plate. In either case, the operation of the tube would be materially affected. The plate structure, as shown in Figure 12-21, is a solid block of copper in which the resonant cavities are cylindrical holes. A narrow slot opens each cavity into the central portion of the tube and divides the inner plate structure into as many segments as there are cavities. Alternate segments are strapped together to put the cavities in parallel so far as the output is concerned. This makes it possible to take the output from a pickup loop placed inside any one of the cavities.

Since the outer conductor of the output coaxial line is connected to the shell of the magnetron, which is a part of the plate structure, there cannot be a high positive DC voltage applied to the plate. The same result is realized, however, whenever a negative voltage is applied to the cathode. The magnets, which are made of a special alloy called Alnico, are unusually strong. Protect them from sharp blows of metallic substances or any hard material, since such blows will cause them to lose magnetism.

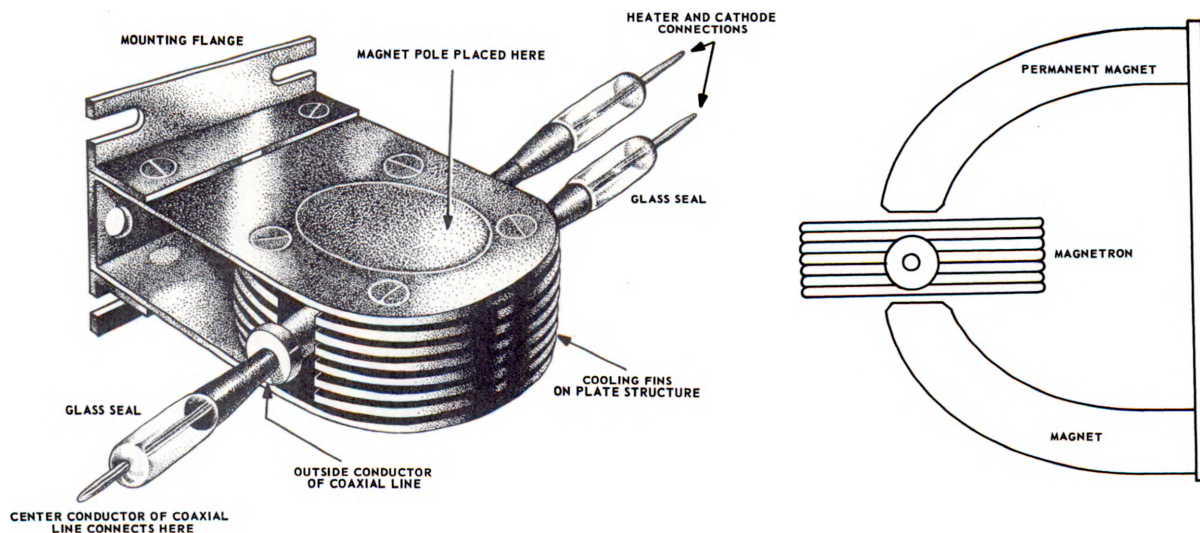


Figure 12-20. Typical 10-cm Magnetron

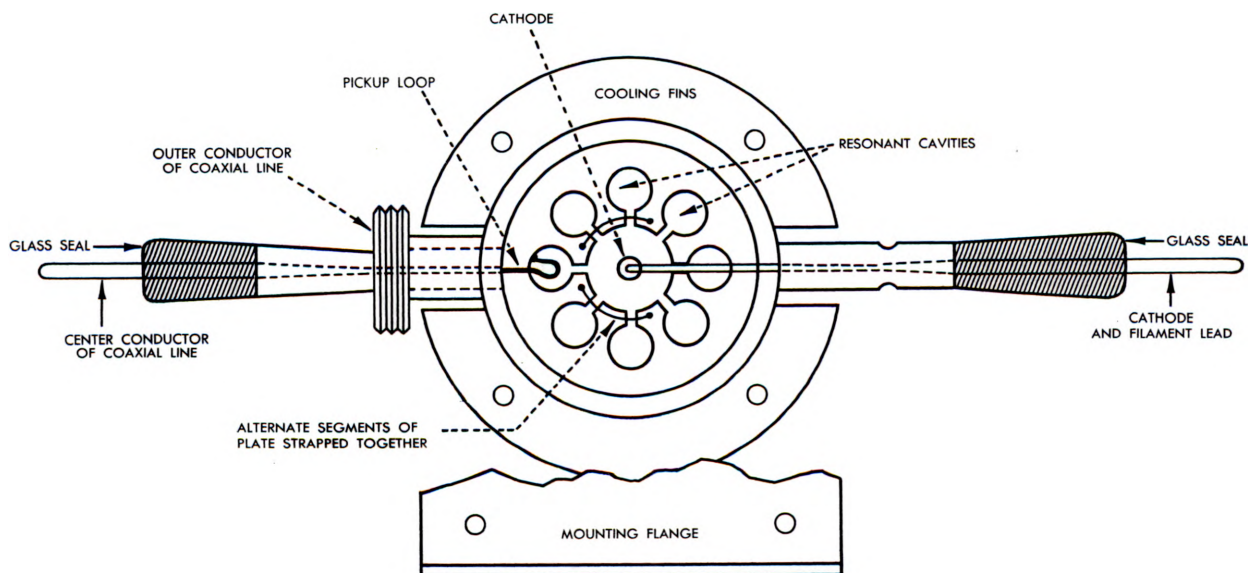


Figure 12-21. Cutaway View of 10-cm Magnetron

### Operation of the Magnetron Oscillator

The theory of operation of the magnetron is based on the motion of electrons in combined electric and magnetic fields. The following laws govern this motion.

The direction of an electric field is from the positive to the negative voltage source. The law governing the motion of an electron in an electric field states that the force exerted by an electric field on an electron is proportional to the strength of the field, and the direction of the force is op-

posite to that of the field. In other words, electrons tend to move from a point of low potential toward one of high potential. Mathematically, this law is stated by the vector equation,

$$F_e = 1.6 \times 10^{-19} E \quad (1)$$

where  $E$  is the electric field intensity in volts per centimeter and  $F_e$  the force in dynes.

The law of motion of an electron in a magnetic field states that the force exerted on an electron in a magnetic field is at right angles to both the field and the path of the electron; the



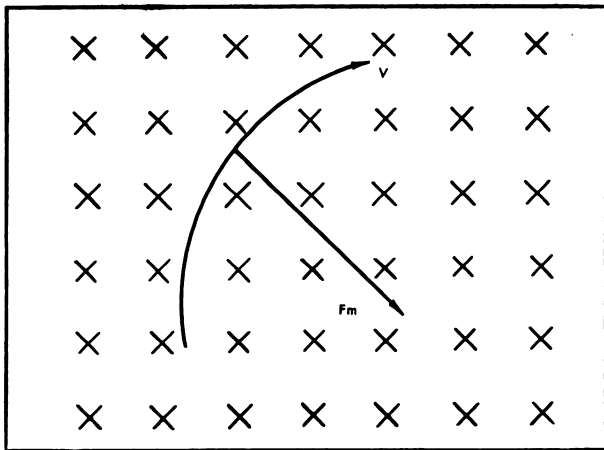


Figure 12-22. Force Acting on an Electron Moving in a Magnetic Field

direction of the force is such that the electron trajectories are clockwise when viewed in the direction of the magnetic field as in Figure 12-22. Mathematically, the law governing the magnitude of a force in dynes is expressed as,

$$F_m = 1.6 \times 10^{-8} V_p B \quad (2)$$

where  $V_p$  is the component of the velocity in centimeters per second of the electron perpendicular to the direction of the magnetic field, and  $B$  the magnetic flux density in gauss.

In regions of uniform magnetic field the electrons move in circles of radius given by the equation,

$$R = 5.68 \times 10^{-8} \frac{V_p}{B} \text{ centimeters} \quad (3)$$

with a period of

$$T = \frac{35.7 \times 10^{-8}}{B} \text{ seconds} \quad (4)$$

The path that an electron takes when traveling normally with respect to the magnetic field depends upon the initial velocity of the electron. Generally, however, the paths of electrons in a plane normal to the magnetic fields are cycloids (figures generated by a point on the radius of a wheel rolling on a smooth plane). Different paths are obtained by different points along the radius of the wheel. For example, the center point of a wheel generates a straight line. A point on the circumference generates a typical cycloid as shown in Figure 12-23. Points lying beyond the radius generate cycloids also, but these are closed loops.

### Paths Generated by Electrons

All paths generated by the rolling wheel may also be generated by electrons traveling at different velocities. This is understandable if you consider that the uniform magnetic field is directed inward perpendicular to this page, and that the electric field is directed downward from the top to the bottom of the page. If an electron is projected horizontally from left to right, there will be two forces acting on it — a magnetic force downward and an electric force upward. If these forces are equal, the path of the electron will be a straight line. Equating equation (1) and (2) gives the velocity required to produce a straight line path. The result is expressed by the equation,

$$V_p = V_s = 10^8 \frac{E}{B} \quad (5)$$

where the velocity to produce a straight path is  $V_s$ .

For all other velocities, the electron path will be curved as shown in Figure 12-24. All electrons traveling with velocities between zero and  $V_s$  take a path similar to that shown in curve b. The qualitative explanation of this path is the following: Because the initial velocity of the electron is less than  $V_s$ , there is a smaller magnetic force acting on it. Hence, the electron curves upward at first. But as it moves, it gains kinetic energy from the electric field with a resultant increase in velocity. The magnetic force thus increases and after a time the electron path is bent downward where the electron begins to lose velocity and finally returns to its original state.

Curve c, Figure 12-24, shows the path that an electron takes when it starts with zero velocity. At the start the magnetic field exerts no force on the electron as no magnetic lines of force are being cut. The electric field, however, moves the electron directly upward to a point where the magnetic field acts on it and deflects the electron to the right. As the electron moves upward, it gains velocity. The magnetic force increases, and eventually becomes large enough to bend the trajectory back until the velocity of the electron again becomes zero. The electron now has the same potential in the electric field as when it started.

An electron traveling from right to left (mathematically, a negative velocity) is at first pushed upward by the magnetic and electric fields as

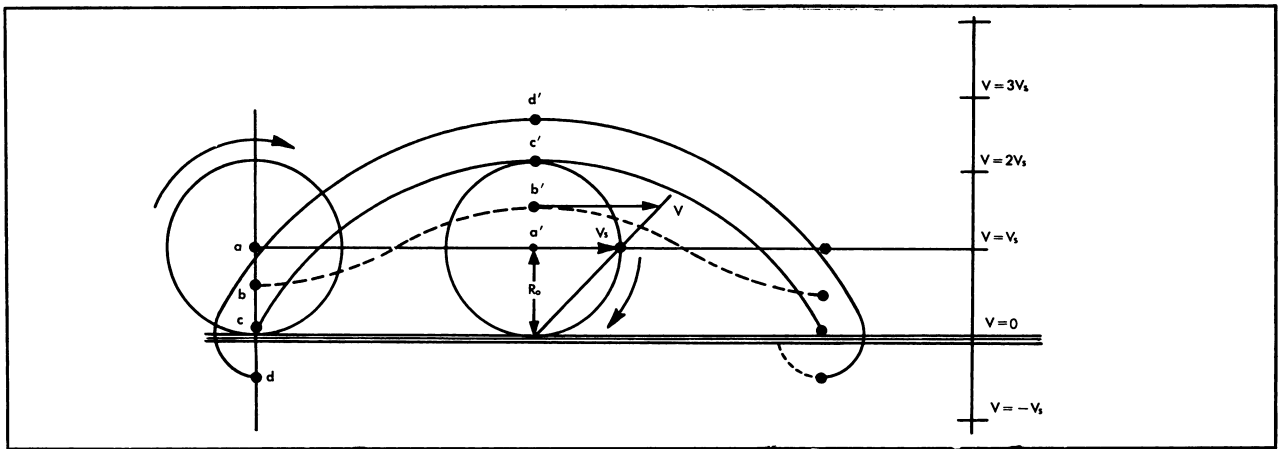


Figure 12-23. Generation of Cycloid

you can see in curve *d*. Its initial radius of curvature is small. Soon it goes straight upward, gaining velocity rapidly. As the effect of the magnetic force increases steadily, the electron is pushed to the right and finally downward. The path is bent back and the electron finally returns to its initial condition.

The time for each cycle and the horizontal distance covered by each electron in one cycle is *the same* for each of the cases given.

All the paths illustrated in Figure 12-24 are cycloids. It follows that electron *a* corresponds to the center of a wheel rolling on a smooth plane. Electrons *b*, *c*, and *d* correspond to a point on the radius, to a point on the circumference, and to a point outside the wheel, respectively. The period of the wheel is identical to the period of an electron in the magnetic field alone, according to equation (4).

### Source of Energy in Magnetron Oscillators

In an electric field alone, a free electron, which always tends to move parallel to the field, gains a large amount of kinetic energy. In combined electric and magnetic fields the electron, which on the average moves at right angles to the field, does not change its average kinetic energy. The magnetic field always exerts a force that tends to change the direction of motion of the electron.

In a magnetron oscillator there are two components in the electric field — a steady component furnished by the DC source and an alternating component furnished by the oscillations of the tank circuit. In a properly designed magnetron, the electrons in transit absorb energy from the DC component of the field and deliver it to the alternating component. The function of the magnetic field, in many respects, is the same

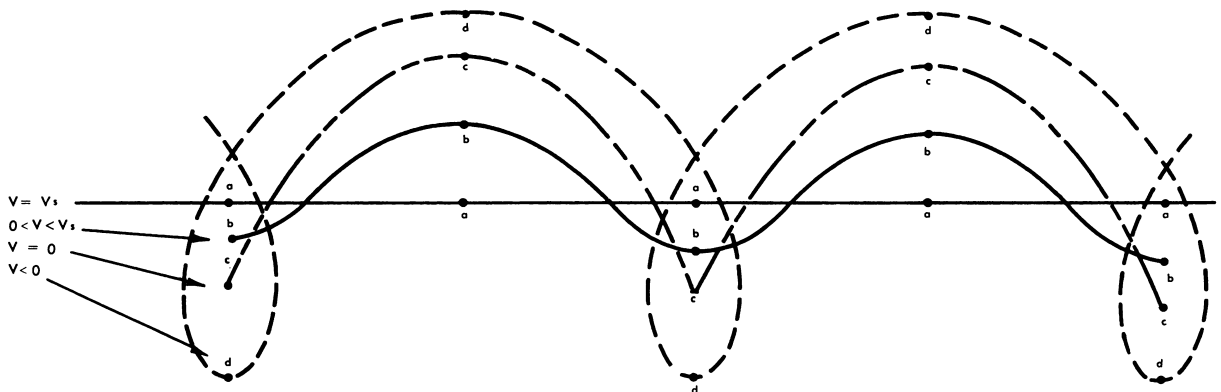


Figure 12-24. Electron Orbits

as that of the positive feedback to the grid of the triode oscillator.

### The Split-Anode Transit-Time Magnetron

The two-segmented-plate cylindrical-anode magnetron is a practical oscillator from which the multi-anode magnetron using resonant cavities as tank circuits was developed (Figure 12-25). In it the electric field is the result of the

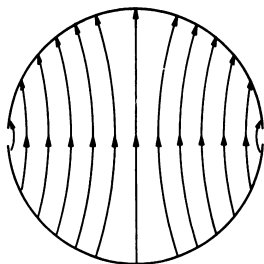


Figure 12-25. Electric Field Between Plate Segments

difference of potential between the two plate segments. In addition to this field, there is a steady field (not shown), which is caused by the potential difference between the cathode and the two anodes. Figure 12-26 shows the ideal motion of an electron that delivers energy to the source on alternating cycles between the magnetron plates. Note that the electron trajectory is approximately cycloidal.

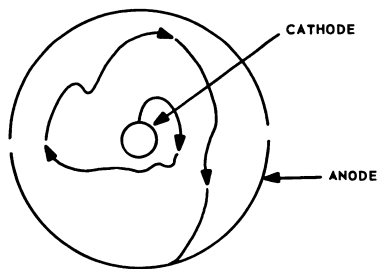


Figure 12-26. Path of an Electron Delivering Energy to the Alternating Field

As you can see, the electric field between plate segments may either increase or decrease the angular velocity of an electron, depending upon the angular position of the electron. If an alternating voltage having a period equal to the time taken for an electron to move once around the

filament is impressed between the anode segments, the direction of the field reverses twice in each revolution of the electron about the filament. As a result, the electrons that pass the gaps at the instants in which the field is maximum will experience continuous angular acceleration or deceleration. Electrons that are accelerated gain energy from the source of alternating voltage. The electrons that absorb energy are speeded up and the component of force toward the filament also increases. As a result, these electrons re-enter the cathode after one excursion.

Electrons that deliver energy, on the other hand, are decelerated. This decrease in angular velocity reduces the average force toward the cathode, and causes electrons to drift toward the plate as shown in Figure 12-26. Experiments have proved that electrons which give up energy make between 80 and 100 excursions before reaching the plate. Since the electrons that deliver energy make so many more radial oscillations before striking the plate than the ones that absorb energy, the source of alternating voltage gains energy. Sustained oscillations will take place whenever the oscillatory circuit has sufficiently low dissipation, such as a Lecher line or a resonant cavity.

### The Multi-Anode Transit-Time Magnetron

**PLANE FORM.** Since the exact theory of the cylindrical multi-anode magnetron is more difficult to understand than that of the plane form, the plane form is discussed first. The ideas presented for the plane form may be carried over in the discussion of the more practical cylindrical multi-anode magnetron.

To understand the operation of the plane form magnetron, consider a magnetron oscillator that is composed of a continuous plane cathode and a segmented anode of which the alternate sections are connected to the opposite sides of its tank circuit. In Figure 12-27, which shows this magnetron, note that the alternating electric field is sketched in the space between the cathode and anode at the instant when alternate anode segments are at their maximum positive and negative values.

Important considerations in the operation of this magnetron are: (1) which electrons tend to sustain oscillations, (2) which electrons tend to

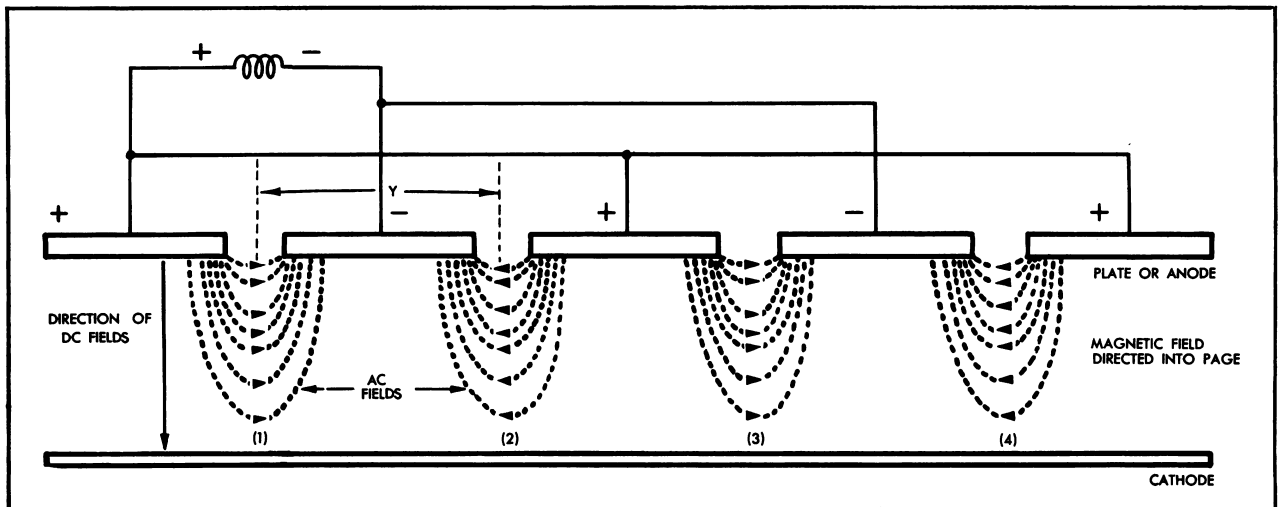


**Figure 12-27. AC Electric Field in Plane Magnetron with Multi-Anodes with Maximum Potential Between Adjacent Anode Segments**

absorb energy, and (3) the probable electron paths. In this connection consider an electron in a uniform electric field directed as shown in region (1) in Figure 12-27. The direction of the resultant field is obtained by adding the AC and DC fields vectorially at this point. (The DC field, which is not shown, is directed downward.) The approximate direction of the electric field and the path of the electron are shown in Figure 12-28A.

As shown in this figure, the electron makes cycloidal paths. Since the general progression of a cycloid is always at right angles to the resultant electric field, these cycloids progress upward and to the right with the magnetic field directed into the page. On the other hand, an electron in a uniform field whose direction is the same as the resultant shown for region (2) in Figure 12-27 would progress downward and to the right as shown in Figure 12-28B. The direction of the uniform electric field is also indicated.

Since the electron of diagram A moves, on the average, from cathode to anode, it must absorb energy from the source of the steady field. This same electron, however, tends to deliver energy to the alternating field since it moves in a direction that is opposed by the field of the tank circuit. Because the average velocity  $V$  of an electron does not change as it progresses toward the anode, it is a very efficient means for converting energy from the DC source into energy of oscillation for the tank circuit. By the same

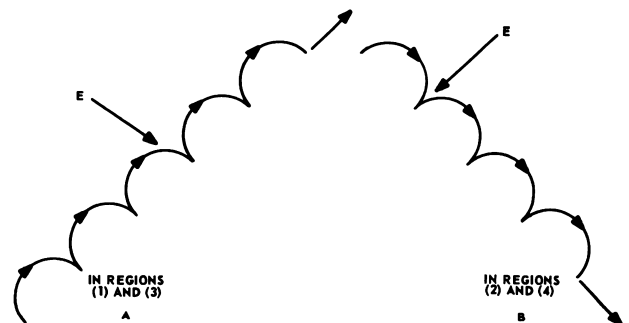


**Figure 12-28. Electron Paths**

reasoning, the electron in region (2) (Figure 12-27) tends to absorb energy from the alternating field and to transfer it to the steady field.

Whenever oscillation is to be maintained, the electron following the path in diagram A must be made to continue its path to the anode or plate. The other electrons — the ones that absorb energy — must be removed from the field quickly. An electron following the path in diagram B, starting from the cathode with a small or zero velocity, will strike the cathode before completing its first loop. Had the fields of regions (1) and (2) been static, the electron of diagram A would tend to turn downward as it approached region (2). Since the fields are alternating, this will not necessarily be the case. In fact, if the left-to-right velocity of the electron of diagram B is such that it arrives in region (2)

in exactly one-half cycle, it will encounter there exactly the same kind of field that it encountered in region (1) one-half cycle earlier.

Thus, if this electron, shown in the magnetron in Figure 12-27, continues to progress distance Y from left to right during each half-cycle, it will go all the way to the anode, following a path similar to that of an electron in diagram A, Figure 12-28. This condition is necessary for oscillation. Oscillation is sustained since the electron of diagram A contributes energy to the tank circuit for a much longer time than the electron of diagram B takes energy away from the tank circuit.

The following analyzes the factors affecting the frequency of oscillation in a plane magnetron. The average left-to-right velocity  $V_x$  of the electron depends only on the average downward component of the electric field. The DC field is the average electric field. Since an electron should progress horizontally a distance of Y centimeters in half a period ( $T/2$ ), it follows that

$$(V_x) \frac{T}{2} = Y \quad (6)$$

$$\text{and} \quad f = \frac{1}{T} = \frac{V_x}{2Y} = \frac{(10^6 E)}{B} \frac{1}{2Y} \quad (7)$$

where  $V_x = \frac{10^6 E}{B}$  from equation (5) and E is the DC component of the electric field.

These conditions clearly indicate that the plane magnetron is a transit-time oscillator, for an electron is allowed a certain definite time to travel from cathode to anode. Thus, in addition to the usual condition of resonance,  $f = \frac{1}{2\pi \sqrt{LC}}$ , there is equation (7) to be fulfilled also. When this is done you may rewrite equation (7) as

$$\frac{f}{(E/B)} = \text{constant} \quad (8)$$

The question arises whether you could obtain greater efficiency if, in addition to adjusting the ratio  $(E/B)$  to the frequency as required by equation (8) or (7), you also made the time for one cycloidal loop equal to the period of the tank circuit. When both of these conditions are filled, the lengths of the cycloidal loops become equal to  $2Y$ , or the length of two anode segments, as illustrated in Figure 12-29. It develops, however, that the efficiency is not increased by doing this.

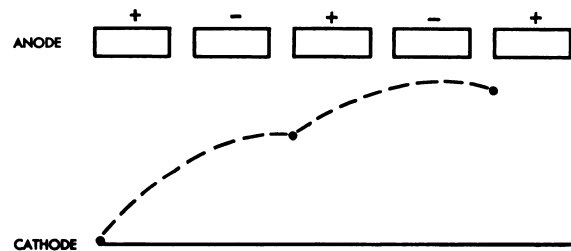


Figure 12-29. Electron Loops

This is why the efficiency is not increased. Since an electron returns to zero kinetic energy at each cusp, the efficiency of conversion from energy to tank-circuit energy is 100% from cusp to cusp. A cusp is a point where two curves or arcs meet. The only DC energy that is not converted into tank-circuit energy by electrons which reach the anode is that which appears as the kinetic energy which arrives at the anode. As this kinetic energy is all acquired after the last cusp, the closer the last cusp is to the anode, the less will be the energy lost relative to the total available DC energy. Therefore, the smaller the cycloids, the smaller will be the fraction of energy lost as kinetic energy at the plate. For these reasons efficiency is not increased by making the looping frequency equal to the tank-circuit frequency.

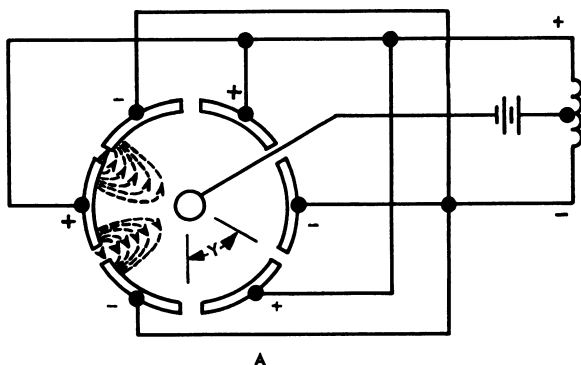
To increase the efficiency in a magnetron, it is necessary to make the cycloids smaller. As can be proved from the properties of a rolling wheel, the size of the orbits depends on the radius of the wheel R, and the equation,

$$R = 5.68 \frac{E}{B^2} = 5.68 \frac{(E)}{B} \frac{(1)}{B} \quad (9)$$

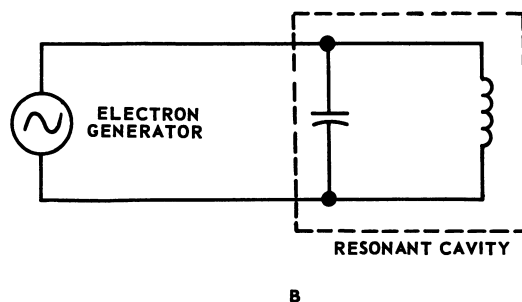
Thus, it follows that for smaller cycloids the value of B must be increased for a given value of  $(E/B)$ . As previously stated the ratio  $(E/B)$  is fixed for a given frequency, as this is one of the conditions for oscillation indicated in equation (8).

In practice it is possible for the value of B to be 1.5 to 3 times the value which makes the looping frequency equal to the tank-circuit frequency. Not only does increasing B increase the efficiency, but it also increases the power input since E must be increased in direct proportion to B in order to keep  $E/B$  constant.

**CYLINDRICAL OR CAVITY FORM.** The cylindrical or cavity form magnetron is analogous to



(A) Simplified Tank Circuit with Six-Anode Magnetron, Showing the Alternating Electric Field.



(B) The Equivalent AC Circuit in Which the Generator Corresponds to the Electrons Between Anode Segments.

Figure 12-30. Six-Anode Magnetron

the plane magnetron rolled up into a circle. The number of segments (anodes) which may be used in it varies widely. As many as 32 have been tried on certain occasions. Here, however, for simplicity the discussion considers only a magnetron with six segments. Figure 12-30A shows the circuit of a six-anode magnetron and the approximate AC field between the segments. Figure 12-30B is its equivalent circuit.

To qualitatively understand the performance of the cavity magnetron, assume that the DC field is uniform and that its magnitude is  $E_b/(b-a)$ , where  $E_b$  is the DC supply voltage,  $b$  the anode radius in centimeters, and  $a$  the cathode radius. Since the AC field is much narrower near the cathode than near the anode in the cylindrical magnetron (shown in Figure 12-30A), the linear distance (the distance which the electron must progress around the cathode per anode segment) is less near the cathode than near the anode. If the theory of the plane magnetron is to be applicable to the cylindrical magnetron, it will be necessary to choose an effective, or average, value of  $Y$ . This value is one-sixth of the circumference of the circle halfway between the cathode and the anode. These assumptions about  $Y$  and the electric field lead to the same resonance condition as before, equation (8). In addition, as with the plane magnetron, it is inefficient to fulfill the resonance condition for B alone.

In developing the discussion of cylindrical or cavity magnetrons, it is desirable to see how some simpler systems of several resonant cavities work. The buncher and catcher cavities of a nonreflex

klystron and a single velocity-modulated electron beam are examples. The theory of buncher and catcher cavities is explained more in detail later in this chapter. Refer to Velocity-Modulated Tubes and the Traveling-Wave Tube for more information on the theory of bunching of electrons.

Figure 12-31 depicts a two-cavity system consisting of two resonant cavities tuned to the same frequency with an electron beam passing through both sets of cavity grids. Assume that the first cavity is excited at its resonant frequency through the coaxial input line, and that the electron beam is velocity modulated. With a properly adjusted system and the correct electron velocity and modulation level, the beam should be bunched at the grids of the second cavity. This being true, oscillations will occur in the second cavity when a large signal is present. The phase of the second cavity oscillations must be such that the bunched electrons are decelerated as they pass through the cavity grids. In the deceleration of the electron beam, energy is given to the cavity.

At any other phase, oscillation will be unaffected by the electron beam if bunched electrons pass through the cavity grids when both are at the same potential. Also, oscillation at any other phase will transfer energy from the cavity to the beam if the bunched electrons pass through the cavity grids at an accelerated rate. Therefore, the electron beam has the capability of transferring energy from one cavity to another.

If the second cavity is moved toward the first, the electron beam then is underbunched at the



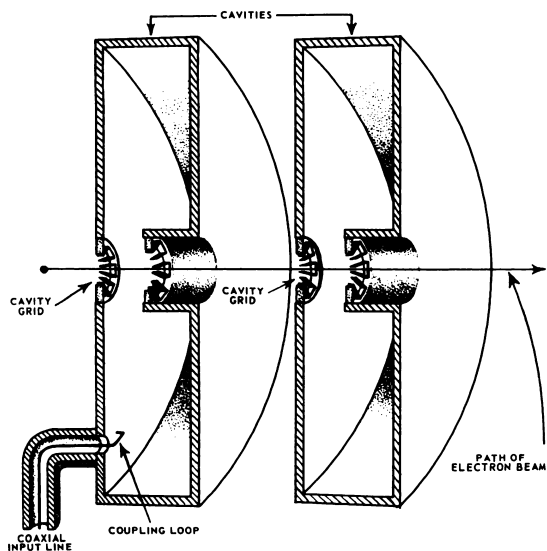


Figure 12-31. Two-Cavity System with Electron Beam Coupling

grids of the second cavity; that is, bunching is not complete. As there is only partial bunching, electron flow is not smooth, but the second cavity is able to absorb energy from the beam. By absorbing energy the phase of oscillation (frequency) in the second cavity automatically adjusts itself in order to slow the greatest possible number of electrons. Those remaining electrons are accelerated as they pass the cavity grids when the second grid is more positive than the first grid. You see, then, that the second cavity slows down over half of the electrons that pass through its grid and accelerates less than half of the electrons. This reduces the average velocity of the electrons and adds to the velocity-modulation of the beam.

In a cavity system that has more than two cavities, it is possible for each cavity, from the second cavity on, to absorb energy from the beam and improve the velocity modulation of the beam for the next cavity. A small portion of the energy comes from the source that feeds the first cavity; most of the energy comes from the electron beam source. Each cavity reduces the average velocity of the electrons in the beam. The system is very sensitive to adjustment, as the energy absorbed by each cavity is dependent upon the relation of the spacing between cavities and the bunching distance.

Let us take a multicavity system, all six cavities tuned to the same frequency, and arrange it as shown in Figure 12-32. Assume that a beam of electrons is influenced so that the beam travels in a closed loop. This loop passes through the associated grids of each cavity. Any irregularity or roughness in the electron beam will cause one or more of the cavities to ring (oscillate). This modulates the beam and energy is delivered to all cavities. The velocity of the electron beam is the source of the microwave energy, and the transfer of energy is a result of the reduction of the average electron velocity. As long as there is a continuous supply of accelerated electrons, while slow electrons are removed from the system, energy will be transferred from the electron beam to all of the cavities. This r-f energy is in the form of electric and magnetic fields. The cavity magnetron operates this same way.

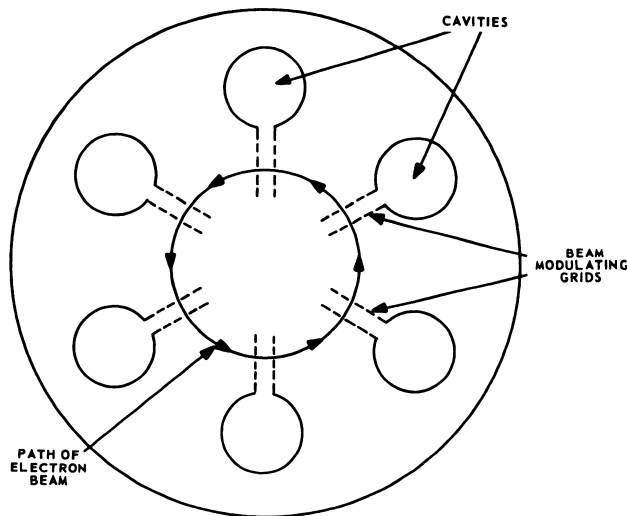


Figure 12-32. Ring System by Resonant Cavities

Figure 12-32 illustrates one of the more common magnetron anodes, a hole-and-slot type. Although six cavities are shown, magnetrons with other numbers and shapes of cavities are not uncommon. Other types of magnetron anodes are shown in Figure 12-33. A shows a slot anode, B a vane anode, and C a rising sun anode. The choice of one anode over another is based largely on design specifications, as all types behave in the same way.

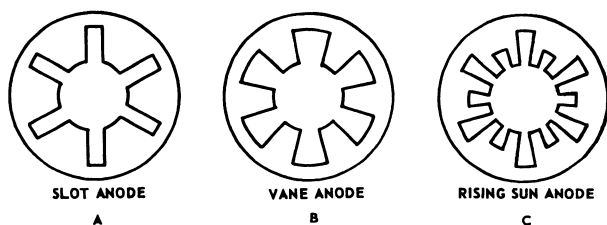


Figure 12-33. Types of Anodes

### MODE OF OSCILLATION OF THE MAGNETRON

The ring system of resonant cavities shown in Figure 12-33, with voltages as applied to the simplified six-anode magnetron shown in Figure 12-30A, appears in Figure 12-34 as six resonant circuits with magnetic coupling between them. While there is usually considerable coupling between inductances that are close together, in actual magnetrons which employ cavities as resonant circuits there is no coupling between adjacent cavities—that is, coupling in a manner characteristic of lumped circuit constants. On the other hand, there can be coupling resulting from radiation of energy from one cavity to another.

As you recall from the theory of coupled circuits, when two circuits tuned to the same frequency are closely coupled, they will resonate not at that frequency but at two different frequencies. The six resonant circuits in Figure 12-33 will resonate at different frequencies separated by a few cycles from each other. A point of interest is whether all these frequencies are

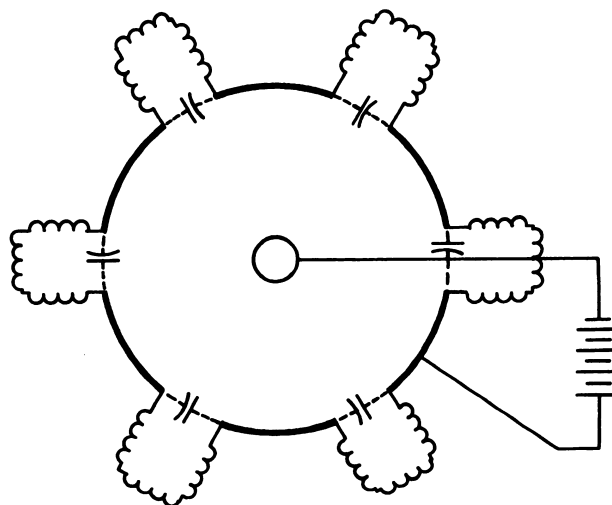


Figure 12-34. Forming Six Magnetically Coupled Resonant Circuits

associated with the same configuration of the AC field in the space between the cathode and anode. As it is, they are not, but instead, for each frequency there is one configuration or *mode* of oscillation. For example, for three different frequencies, there are three configurations as illustrated in Figure 12-35.

In connection with the excitation of mode 3, the average velocity,  $V_a$ , which is equal to  $10^8$  E/B, must be adjusted so that an electron passes two segments per cycle. In mode 2 the ratio of E/B must be higher since the electron must pass three segments per cycle. It follows that this mode requires very high voltage and is difficult to excite strongly.

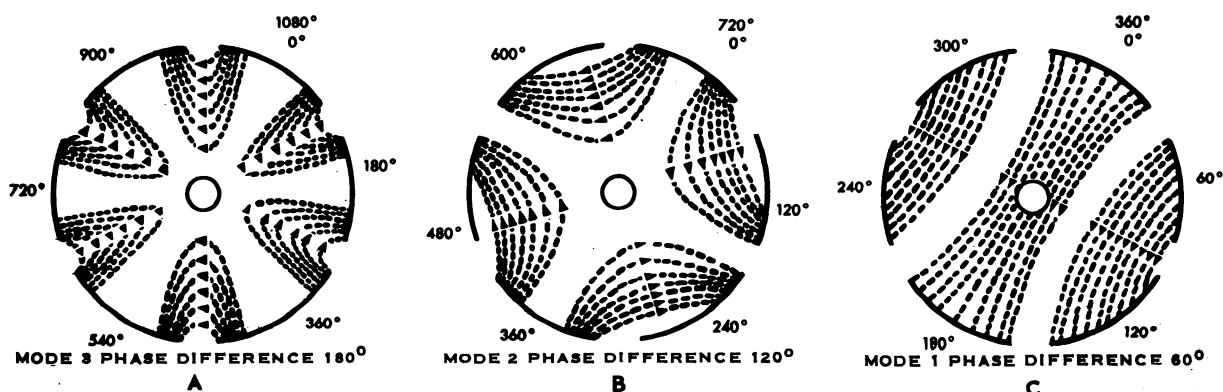


Figure 12-35. Configuration of AC Field Between Adjacent Anode Segments in the Various Modes of Oscillation

The radar magnetron should oscillate only in one mode under operating conditions, since the frequencies of these three modes are generally different, and since the radar receiver can be tuned to only one frequency at a time. If the modes are widely separated in frequency, it will be easier to excite only the one desired. Since mode 3 requires the least voltage, it is the mode generally used.

Another method of describing the modes of operation is as follows. In the case of the six-cavity magnetron, the phase-difference between adjacent cavities may be  $60^\circ$ . Then the sum of the six-phase differences is 1 cycle. When the phase difference is  $120^\circ$ , the sum is 2 cycles; if the phase difference is  $180^\circ$ , the sum is 3 cycles. In like manner, in an 8-cavity magnetron the phase difference between adjacent cavities may be  $45^\circ$ ,  $90^\circ$ ,  $135^\circ$ , or  $180^\circ$ . A phase difference of  $180^\circ$  is the desired mode, and other modes frequently are suppressed by strapping.

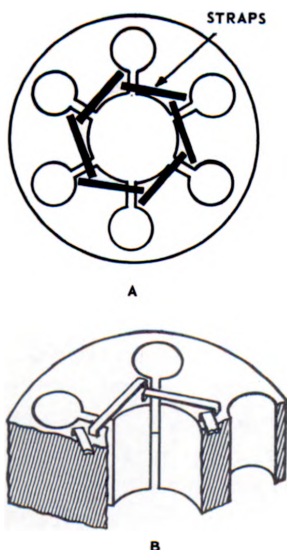


**Figure 12-36. Strapped Anode**

Strapping is illustrated in Figure 12-36 for a six-cavity hole-and-slot anode. Figure 12-36A shows an end view of the strappings and B shows a three-dimensional view. Corresponding points on adjacent cavities are connected by strappings. As the strappings have an electrical length of a half-wavelength,  $180^\circ$  phase difference is assured. Only one lip of each cavity is shown strapped,

but the other lips on the other side of the anode are strapped also. There are other strapping arrangements; all of the strappings may be at one end or a complete set of strappings may be at each end. A magnetron like this is called a strapped magnetron.

### Summary

The following summarizes magnetron theory.

1. By using a segmented anode, you can make the electrons work against an alternating electric field that is crosswise to the steady field. In this way it is possible to get the electrons to go all the way to the anode without much increase in kinetic energy. This results in high power and high efficiency.
2. Oscillations are obtained in mode 3 ( $\pi$ ) when the average crosswise velocity  $V_a$  is such that the electron passes two plates per cycle. The equation for resonance for these oscillations is,

$$f = 10^8 \frac{E}{2BY}$$

In using this equation replace factor 2 by 3 and 6, respectively, for modes 2 and 1.

3. The frequency of looping, which depends on  $B$  alone, is not usually matched to the other frequencies, because doing this causes too low a power input and too low an efficiency.
4. Quantitative calculations show that there are more electrons that give energy to the tank circuit than electrons that remove energy from the tank circuit. Further, it shows that on the average each of the former electrons gives more energy to the tank circuit than each of the latter takes away.
5. Magnetrons in use at present have a wavelength range from about 50 cm down to less than one cm.
6. Efficiencies of magnetrons at moderate power in service at present have values from 30% to 40%, and from 50% to 60% at powers greater than 200 kw.
7. The actual tuned circuit will resonate at several different frequencies, each of which corresponds to a distinct phase difference between anode segments.
8. Magnetrons are readily adaptable to pulsed modulation.

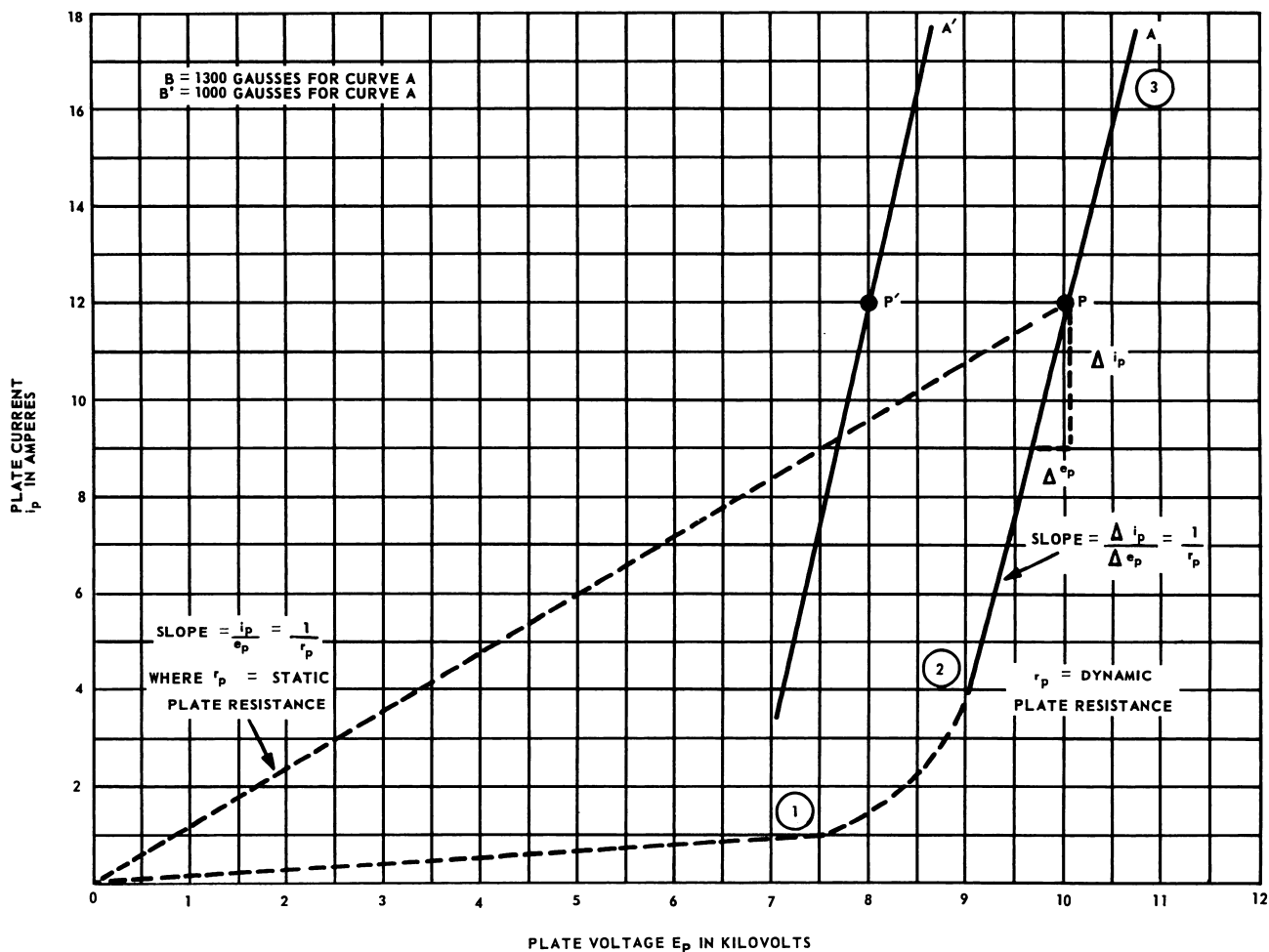


Figure 12-37. Magnetron  $e_p$ - $i_p$  Characteristic Curve

### Plate Current-Plate Voltage Curves of a Magnetron

A useful curve for studying magnetrons is the  $e_p$ - $i_p$  characteristic curve. The curve shown in Figure 12-37 is a typical curve for a 10-cm magnetron operating at about 1300 gauss. Ordinarily, these curves are plotted by holding the magnetron field constant.

These curves are important in that you can get much information by proper interpretation. Keep in mind the following factors about these curves. The resistance between the cathode and anode is nonlinear. The AC and DC resistances are different from each other at any one voltage and vary with the voltage. For example, at point P the static resistance  $r_p$  is the ratio of  $e_p/i_p$ , while the dynamic resistance is the ratio of  $\Delta e_p/\Delta i_p$ . Since the magnetron operates on the

straight line portion of the curve, the reciprocal of the slope of this straight portion, ② to ③, is the dynamic resistance  $r_p$ .

Note that from the origin to point ① on curve A the magnetron current is very small. A negligible coherent r-f energy is produced. Also a considerable amount of random r-f noise is produced. Somewhere between 6 and 8 kilovolts, appreciable oscillation starts. In the region from point ① to point ②, the operation of the magnetron varies greatly, not only from one magnetron to another, but also from one time to another in the same magnetron. Oscillations of more than one wavelength can be detected in the output, indicating that the magnetron is oscillating in more than one mode. The value of  $E/B$  which excites oscillations of higher modes is probably a submultiple of that which fulfills the



resonance condition for these modes, and at the same time is not very far from the resonance condition for the 180° mode.

In the straight-line portion, between points ② and ③, the magnetron oscillates in the stable mode 3. In this region the frequency varies only a few megacycles over this range. In the strapped 10-cm service magnetrons the straight portion continues to the highest current obtainable. Theoretically, if the range is made great enough, mode 2 might be expected to appear. However, this mode is rarely, if ever, encountered in strapped tubes. Unstrapped magnetrons, however, have an additional multimoding region near or about point ③.

Even though the percentage of variation in frequency is small over the straight line portion, the actual change may be greater than the receiver bandwidth. For this reason it is important to maintain the pulse amplitude constant.

The power input is  $e_p i_p$ . Since the straight line portion is fairly steep, the input power rises rapidly with increase in voltage.

The magnetron has a very low efficiency in the region 0—①—②. In the single-mode region ②—③, the efficiency is maximum somewhere in the middle of the straight line portion and decreases slowly toward point ③. In this region you may safely assume that the efficiency is constant and equal to 30–40% over the entire region. It may seem surprising in view of the resonance condition specified earlier, that a single-mode oscillation is excited over such a wide range of voltage (region ②—③). This is explained by the fact that the point of resonance is not a sharp one. The point of resonance on the characteristic curve is not the only point at which oscillation is excited, but is the point where the *efficiency* of operation is a maximum.

The power output increases approximately as the power input in region ②—③ since the efficiency is practically constant within this region.

Curve A' shows the effect of reducing the magnetic flux density to 1000 gauss. To a close approximation, the new operating point P' will occur at a voltage 25% less than that of P. The current will be the same. As is evident, the dynamic resistance will be essentially unchanged. If B were increased 25%, the curve would be shifted to the right of curve A by about the

same amount. The general properties of  $i_p$ - $e_p$  curves are not altered except that both the input power and efficiency increases in the linear portion as B is increased. The increase in power input is the result of the higher voltage necessary to keep E/B constant. The efficiency increases because the kinetic energy with which the electrons strike the anode remains roughly constant (from equation (5)), while the total energy available increases with the voltage.

Table 12-1 summarizes the characteristics of the magnetron:

| Region               | Mode                | Power Input       | Power Output      | Efficiency                     |
|----------------------|---------------------|-------------------|-------------------|--------------------------------|
| 0-①                  | Noise               | Very Small        | Zero              | Zero                           |
| ①-②                  | More than one       | Small             | Very Small        | Very Small                     |
| ②-③                  | Mode 3              | Increases rapidly | Increases rapidly | High and fairly constant (40%) |
| Beyond ③ Un-strapped | Double-mode         | Large             | Moderate          | Low                            |
| Beyond ③ Strapped    | Continues same mode | Increasing        | Large             | Moderate and decreasing        |

Table 12-1. Summary of Magnetron Characteristics

### Modulating the Magnetron Transmitter

The modulator for the magnetron serves a double purpose—it determines the waveform of the output pulse, and it stores energy between pulses and releases it through the magnetron during the pulse.

There are two circuits that make it possible for the modulator to store and release energy. These are the parallel charging circuit and the series charging circuit.

As illustrated in Figure 12-38A, the source of energy in the parallel charging circuit is connected at all times to the storage element through the charging impedance  $Z_{ch}$ . During the output

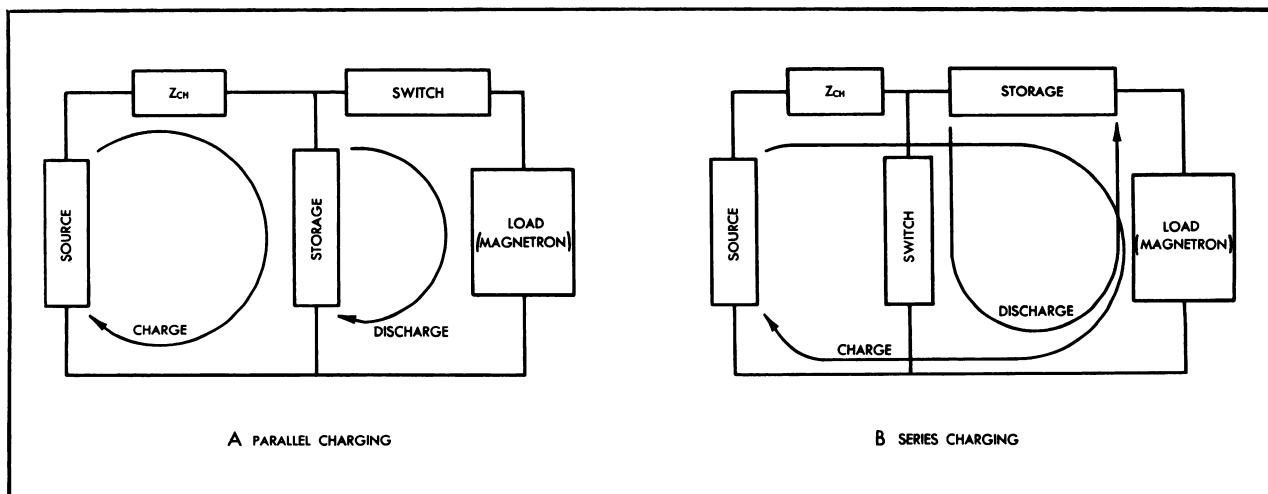


Figure 12-38. Parallel and Series Circuits for Storage and Release of Energy

pulse the closed switch applies the voltage from the storage element to the load.

A disadvantage of this circuit is that it is not possible to prevent grounding of all three of the circuit elements—the source, the switch, and the load.

Because of the disadvantages of the parallel charging circuit, the series charging circuit is generally employed. In this circuit the storage element is charged through the charging impedance and the load. It is discharged through the load when the switch is closed. From an examination of this circuit, you can see that closing the switch also completes a circuit containing the source and  $Z_{ch}$ . This has little consequence, however, since the switch is only closed for about a microsecond at a time, and  $Z_{ch}$  is large enough to prevent any damage.

It is well to find out the components of some of the modulator circuit elements which are shown as blocks in Figure 12-38. The source from which the storage element is charged may be either a DC voltage or an AC voltage. A magnetron requires a voltage of 10 to 15 kilovolts or higher for proper operation. In some types of equipment a transformer is used between the storage element and the magnetron to step up the voltage, while in many others a direct connection is made between the storage element and the magnetron. Actual charging circuits are discussed in a later section dealing with line-pulsing modulators. In these circuits the charging impedance may be a high resistance or a

high inductance, or both. Its purpose is to control the charging time for the storage element and to prevent short-circuiting of the source during the pulse.

Switches for controlling the charging time for storage are of two general types—a soft switch or a hard tube. A soft switch is one in which conduction takes place through a gas. Soft switches include gas thyatron, mercury thyatron, fixed trigger gap, and rotary spark gap switches. A hard-tube switch is a high-vacuum-tube switch. Up to a few years ago, radar equipment used the hard-tube, spark gap, and thyatron switches. The present trend is to use hydrogen thyatron tubes as a switch because of the stability requirements of the Moving Target Indication system.

In the fixed trigger gap switch, Figure 12-39, breakdown occurs at the desired instant when a trigger pulse is applied to electrode C. This, in turn, causes breakdown from C to B resulting

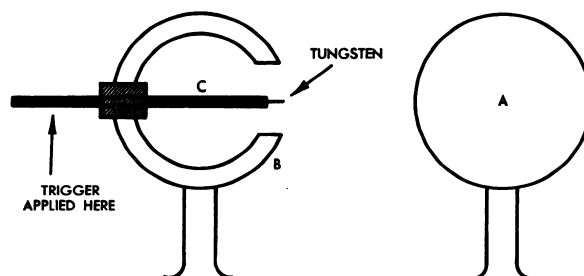


Figure 12-39. Fixed Trigger Gap



in the formation of a sufficient number of ions to cause breakdown between the high-voltage electrodes A and B. One difficulty with this switch is that the trigger wire C corrodes and is consumed in too short a time to make the circuit general in use.

In the rotary spark gap shown in Figure 12-40, there are four electrodes on the rotating wheel, which is grounded. The electrodes pass between two pairs of fixed electrodes when the switch is in the 800-cycle position, and between one pair when it is in the 400-cycle position.

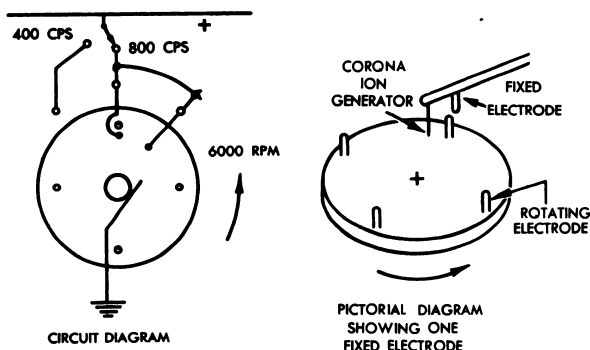


Figure 12-40. Rotary Spark Gap

Each time one of the moving electrodes passes a pair of fixed electrodes, there is a breakdown of voltage and the switch closes for a brief period of time. In operation there is some irregularity in the intervals between breakdown, which may cause the period to vary as much as 50 microseconds. But in a self-synchronous radar set this presents no difficulty. In fact, there is some advantage to the irregularity since it causes pulses from other radar sets to move on the scope, and it also lessens the likelihood of confusion due to echoes occurring on the second trace after the transmitted pulse.

Each type of switch mentioned has certain disadvantages; none is perfect for the job. Therefore, you may expect to find any or all of them in use in various sets.

The storage element can be either a capacitor or a pulse-forming network. A capacitor is most often used with the hard-tube modulator in which the sole purpose is to store energy. In this type of modulator circuit, it is necessary that the time constant be such that the capacitor does not discharge appreciably during the pulse for,

as you can see in Figure 12-37, the magnetron current and, in turn, the magnetron output drop very rapidly when the voltage decreases.

The pulse-forming network in the storage element serves the double purpose of storing energy and of shaping the output waveform. In operation it is quite similar to the artificial transmission line; that is, it furnishes a steady output voltage for a certain length of time at which time its voltage drops to zero. It is different, however, in that instead of being made up of a number of identical sections, each section is different. It has the advantage of forming more nearly rectangular pulses than an artificial transmission line with a similar number of sections.

Although the theory of operation of the pulse line shown in Figure 12-41 during discharge is somewhat complex, it may be summed up as follows:  $C_4$  stores most of the energy for the pulse while the components  $C_1$ ,  $C_2$ ,  $L_1$ ,  $L_2$ , and  $L_3$ , which are in series with it, mainly shape the pulse. During the first part of the pulse (when the charge on  $C_4$  is nearly equal to the supply voltage  $E_b$ ) the series components offer a high impedance and cause the voltage across  $Z_c$  (the load) to jump to  $E_b/2$ . As the pulse continues,  $C_4$  discharges but the impedance of the series components also drops. This causes the voltage across  $Z_c$  to remain at  $E_b/2$ . Later when the charge on  $C_4$  is less than  $E_b/2$ , the inductances maintain the level of current flow. This continues until the capacitor is discharged and the magnetic fields are collapsed. At this time the voltage drops rather rapidly to zero.

The components  $C_3$  and  $R_1$  make the leading edge of the pulse steeper. There being no inductance in this branch of the network, the current can rise immediately to its full value. The

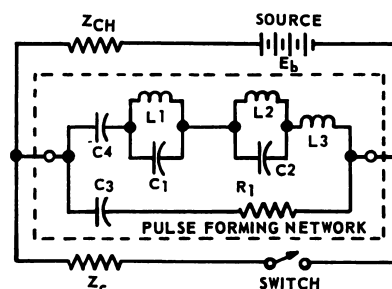


Figure 12-41. Pulse-Forming Network as Storage Element

pulse that appears at the load is essentially a rectangular pulse of the desired duration and equal in amplitude to  $E_b/2$ . The values of the components determine the duration of the pulse and the characteristic impedance of the line. The characteristic impedance of the line must be matched to that of the load. For this reason, even though the line is charged to  $E_b$ , the voltage across  $Z_c$  is only  $E_b/2$ .

Study Figure 12-42 to get a comparison of the output pulse and the discharge curve of  $C_4$ . The time of rise of the voltage is about one-tenth of the pulse duration and the time of fall a like amount of time. This type of pulse network is usually sealed in a metal box filled with oil which acts as an insulator. For that reason they are sometimes called potted networks. It is impossible to replace components in the line. Some lines using only a few elements are not potted. Generally, these do not give as nearly rectangular an output pulse as the potted networks, due to the smaller number of sections.

In the circuits discussed thus far the load can be either a magnetron and its associated circuits or the primary of a pulse transformer that has its secondary connected to the magnetron. In addition, a pulse transformer can be used in the amplifier stages preceding the hard-tube modulator. When this is done, all tubes are cut off between pulses. The interstage transformer is a low-voltage and low-turns-ratio device. Its only purpose is polarity inversion. The use of a high-voltage step-up transformer to furnish voltage for the magnetron makes possible the use of lower voltages in the pulse-forming network and thus simplifies the insulating problem.

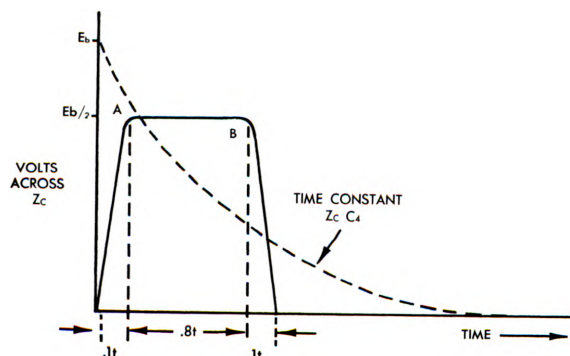


Figure 12-42. Comparison of Pulse from Pulse Line with Exponential Discharge of Capacitor

Another advantage in using a transformer with two secondaries, as shown in Figure 12-43, is that it avoids having high voltage in the magnetron filament transformer. A third need for the high-voltage pulse transformer is matching impedances, particularly where the radar antenna is remote from the set itself. Since the use of long transmission lines between the magnetron and the antenna make the magnetron's operation somewhat erratic, it is desirable to have the magnetron near the antenna. This necessitates the use of long lines to carry the modulator pulse to the magnetron. In order to be able to use a 50-ohm line for this purpose and still get a match of impedance with the approximately 1000 ohms of the magnetron, it is necessary to use a  $4\frac{1}{2}:1$  step-up transformer at the magnetron end.

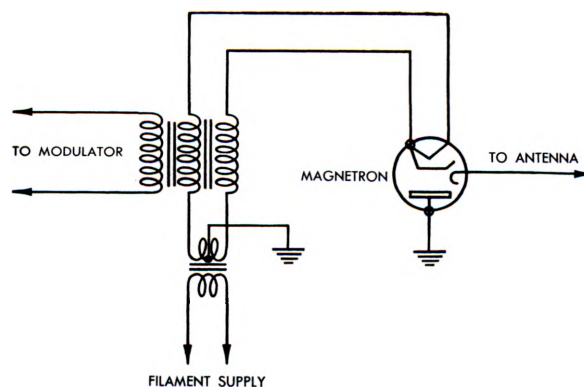


Figure 12-43. A Transformer with Two Secondaries Makes Possible the Grounding of the Filament Transformer Secondary

The pulse transformer must be specially designed because there are very high frequency components present in a rectangular pulse of such short duration. The core is usually silicon steel and has laminations in the order of 0.003 inch in thickness. In general, a good pulse transformer must have low leakage inductance, low interwinding capacitance, and high primary inductance.

Low leakage inductance is necessary in transformers to preserve the steep leading edge of the pulse. This is achieved by very close coupling between primary and secondary and by using fewer turns on each. Low interwinding capacitance causes the oscillations between it and the

leakage inductance to be of high frequency, making it easier to damp them out. Since capacitance between windings can be decreased by separating the windings farther, you see that a compromise is necessary. In most cases the spacing is rather close and the number of turns low, the primary and secondary consisting of a single layer each, wound one on top of the other on the same leg of the core.

To damp oscillations, a resistor is shunted across the primary or the secondary, or both. The high primary inductance is desirable to induce a high voltage into the secondary, and to insure that there is little change in the secondary voltage if the primary current remains constant. High primary inductance requires many turns, which is in conflict with the requirement for low leakage inductance. Usually the latter is considered more important so the primary inductance is not as high as would be desirable. For this reason the secondary voltage does drop some during the pulse. The maximum allowable drop is about 5%.

### Types of Modulators

There are two types of modulators: the line-pulsing modulator and the driver-hard-tube modulator, Figure 12-44. The line-pulsing modulator stores energy and forms pulses in the same circuit element. This element is usually the pulse-forming network. The driver-hard-tube modulator forms the pulse in the driver and stores the energy in the modulator circuit.

The advantages and limitations of the various types of charging circuits are shown in Figure 12-45.

In circuit 1 of Figure 12-45 the pulse line is represented as  $C_{st}$ . It is charged through  $R_{ch}$  to the value of the DC voltage. During the pulse one-half of the DC voltage is applied to the magnetron (assuming the line impedance is matched to that of the magnetron) for a period of time determined by the line components. This is only 50% efficient. Therefore, it is necessary that the DC voltage be twice the value needed for the magnetron. The time constant  $R_{ch}C_{st}$  must be large in comparison to the pulse length  $d$  but small in comparison to the period of the repetition frequency. This insures that the line will fully charge between pulses but will not discharge appreciably through the source and  $R_{ch}$  during the pulse.

In circuit 2 of Figure 12-45 resistor  $R_{ch}$  is replaced by an inductance  $L_{ch}$ . As  $C_{st}$  charges, the current through  $L_{ch}$  builds up a magnetic field. This field causes current to continue after  $C_{st}$  is charged to  $E_b$ . The charge will reach a peak of approximately  $1.9 E_b$  after which it starts decreasing. If the modulator is triggered at the instant of maximum charge, the voltage applied to the magnetron is about 95% of  $E_b$ . This necessitates that the resonant frequency of  $L_{ch}$  and  $C_{st}$  be one-half the PRF of the radar. The firing time is very critical if you are to take advantage of the maximum voltage.

The difficulty encountered in the preceding circuit is overcome in circuit 3 by adding a diode

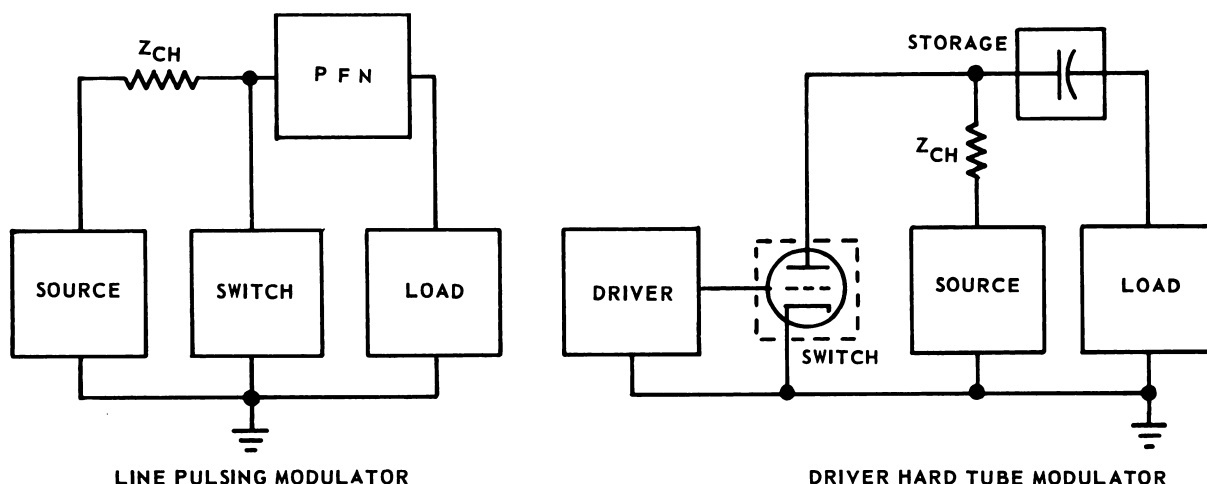


Figure 12-44. Two Types of Modulators

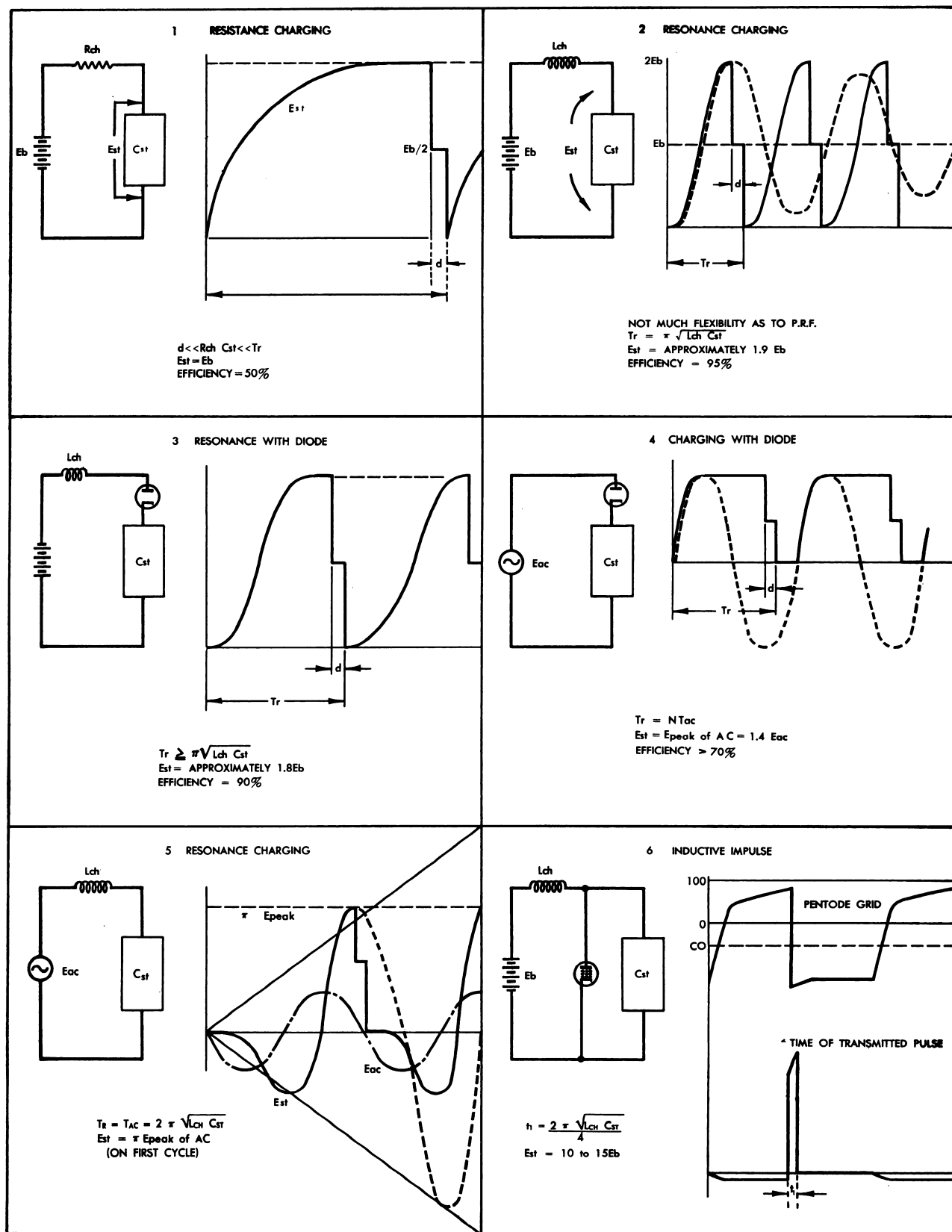


**Figure 12-45. Types of Charging Circuits**

in series with the charging element. This prevents the line from discharging after reaching its peak. The time of firing can occur any time after the peak is reached. Therefore, the frequency at which  $L_{ch}$  and  $C_{st}$  resonate is equal to, or greater than, one-half the PRF. The disadvantage of this method of making the firing time less critical is that the voltage to which the pulse-forming line charges is decreased by the drop across the diode to about  $1.8 E_b$ . This makes the efficiency of this circuit about 90%.

Circuit 4 has the pulse line charging from an AC source. In this circuit the charging line is prevented from discharging by a diode in series with it. In this circuit it is necessary that every trigger pulse come at a time when the diode plate is negative; otherwise, the diode and the switch will short-circuit the AC source. Therefore,  $T_r = N T_{ac}$ . The voltage to which the line charges is almost the peak value of the AC wave so that the efficiency is about 70%.

In circuit 5 the diode is replaced by the inductance  $L_{ch}$ . This inductance, with  $C_{st}$ , forms a resonant circuit that is resonant at the frequency of the AC supply voltage. In this circuit the switch may be closed at any peak of oscillations. Usually, though, it is closed at the first peak. At this time the voltage is approximately  $\pi$  times the peak value of the AC voltage. In addition  $T_r = T_{ac} = 2 \pi \sqrt{L_{ch} C_{st}}$ . If the switch is

closed at the second peak, the voltage will be nearly twice as high. This type of charging gives voltages higher than the source voltage.

The inductive impulse charging circuit in Figure 12-45 provides voltages 10 to 15 times as great as the supply voltage. It is used where the available supply voltage is of the order of 1 kv and operates as follows: The pentode grid is raised above cutoff and into the positive voltage region so that the current through the inductance builds up a magnetic field around it. At time  $t_1$  (the time before the beginning of the transmitter pulse) the pentode is suddenly cut off. The magnetic field in collapsing sends current through the pulse line and charges it to a very high voltage. To take full advantage of this high voltage, the transmitter must be keyed at the instant of maximum voltage. The correct timing is  $t_1 = \frac{1}{4} (2 \pi \sqrt{L_{ch} C_{st}})$ . The value of  $t_1$  is usually 10 to 15 microseconds. This type of charging circuit is used only in conjunction with a master timer.

**LINE-PULSING MODULATOR.** The typical line-pulsing modulator circuit in Figure 12-46 uses an AC resonance-charging circuit. It also has a 115v AC input which is stepped up to 3250v by an input transformer. The resonance-charging system charges the line to approximately 2.8 times the 3250 volts or about 9 kv. The switch is a rotary spark gap which fixes the PRF at 400

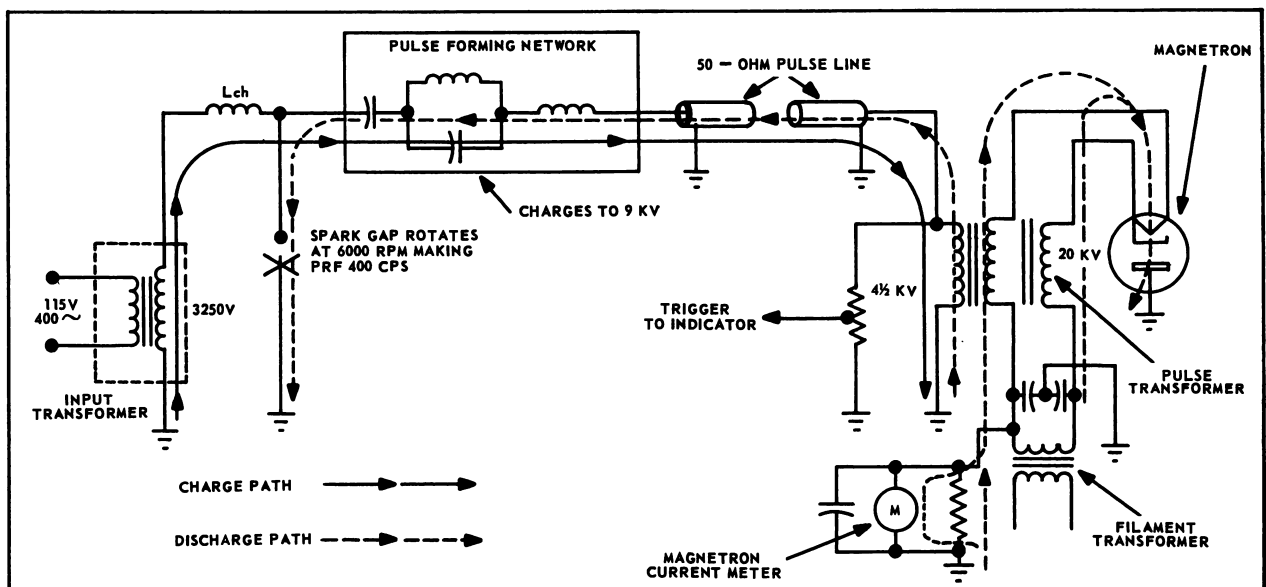


Figure 12-46. Typical Line-Pulsing Modulator



cps. The pulse-forming line is connected through a 50-ohm pulse line to the primary of a pulse transformer. The voltage across the primary is approximately 4.5 kv. This is stepped up to about 20 kv by the transformer.

**THE DRIVER-HARD-TUBE MODULATOR.** The driver-hard-tube modulator, as the name indicates, consists of two parts — the driver and the hard-tube modulator. The hard-tube modulator is the part that stores energy between pulses to be released when the hard tube conducts during the pulse. The driver is the part that shapes the pulse applied to the grid of the modulator.

In the earlier days of radar the hard-tube modulator was the most commonly used circuit, because at that time there was no good switching arrangement available for the line-pulsing modulator.

A simplified circuit of a hard-tube modulator using DC resonance with diode charging is shown in Figure 12-47. Between pulses from the driver, the modulator tube is nonconducting due to a very high negative voltage on its grid. The storage capacitor charges through the path indicated. This path includes  $R_1$ ,  $L$ , and  $R_2$  and the milliammeter in parallel. As  $R_2$  is about 1000 ohms, most of the current flows through the meter. The purpose of  $R_2$  is to maintain a charge path when the meter is switched to other circuits to measure current. Because the meter reads the charging current, it will indicate the average magnetron current since the charge that is lost during the pulse is replaced between pulses.

When the driver raises the grid potential of the modulator, it starts conducting and the capacitor discharges through the path containing

the magnetron and the modulator tube. The charging diode prevents any discharge through that branch. A negligible amount of current will flow through the source,  $R_1$  and  $L$ . When the driver lowers the modulator grid voltage below cutoff, the pulse ends.

The amount of voltage that the modulator applies to the magnetron is determined by a voltage divider circuit consisting of the static plate resistances of the magnetron and the modulator. These resistances have values of approximately 1000 ohms and 100 ohms, respectively.

The voltage that is applied to the magnetron is roughly 90% of the capacitor voltage. It is not a true rectangular pulse, however, because of the distributed capacitances in the leads to the two capacitor plates and to the magnetron. The total distributed capacitance, which is 50 to 120 micromicrofarads, must be charged as the voltage rises (in a negative sense) and discharged as the voltage decreases. The time constant for charging the distributed capacitance, which is determined mainly by the static plate resistance of the modulator tube, is on the order of 0.01 microsecond, a time which is small enough to ignore.

Discharge of the distributed capacitance takes place through  $R_1$ ,  $L$ , and the magnetron itself. The calculation of the discharge time constant is rather complicated, but you can estimate it by ignoring the effect of the inductance and figuring the parallel combination of  $R_1$  and plate resistance of the magnetron as the voltage decreases. For example, an average discharge time constant is about 0.5 microsecond as  $R_1$  and the  $r_p$  of the magnetron are relatively high. This dis-

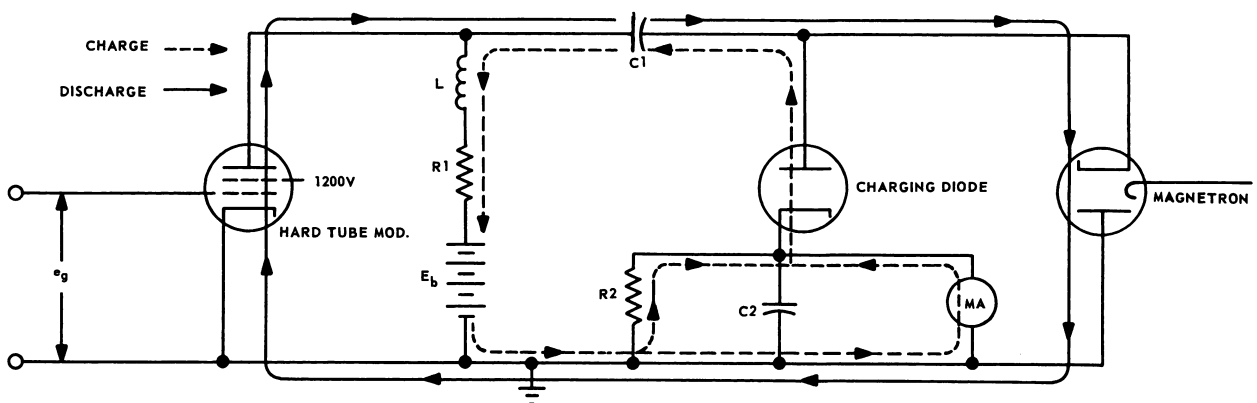


Figure 12-47. Hard-Tube Modulator with Diode Charging



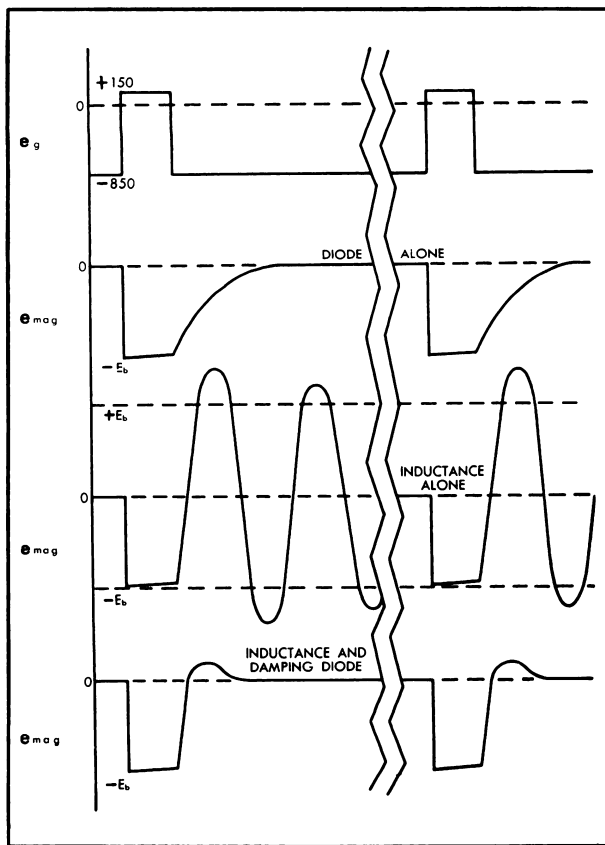


Figure 12-48. Voltage Waveshapes in Hard-Tube Modulator Circuit

charge time causes the voltage to trail off rather slowly as shown in Figure 12-48 and, in turn, produces considerable noise and erratic oscillations for one or two microseconds as the voltage passes through the region below 8 kv. All this makes it impossible to detect a nearby target.

The difficulty is overcome by using a tail-clipping circuit, which consists of an inductance and a diode connected in parallel, to replace the charging diode. In the modulator tail-clipping circuit in Figure 12-49, the inductance ( $L_2$ ) is small, being 3 to 5 mh. In the charge and discharge paths, most of the charging current flows through  $L_2$ . This means that you could leave the diode out of the circuit as far as charging is concerned except for a very important function. The discharge path includes  $L_2$  in parallel with the magnetron. At the beginning of the pulse little current flows through the inductance. Nevertheless, it builds up because the inductance is small. At the end of the pulse the inductance discharges through the distributed capacitance. This causes the voltage to return more rapidly to zero. Another point is that  $L_2$  and the distributed capacitance can be thought of as a resonant circuit. With a small value of  $L_2$  the resonant frequency is such that the period is only two or three microseconds.

The oscillations of this tuned circuit would cause the voltage to swing positive and negative if it were not for the damping diode which conducts when its plate swings positive and ends the oscillations. For the waveshapes with and without the damping diode, see the bottom illustrations in Figure 12-48. Since the current through the inductance has to build up during the pulse, there will be more change in magnetron voltage (faster discharge of  $C_1$ ) than with diode charging alone. The larger the tail clipping inductance, the less the magnetron voltage will change. Already you have seen that in order to return the magnetron voltage to zero quickly

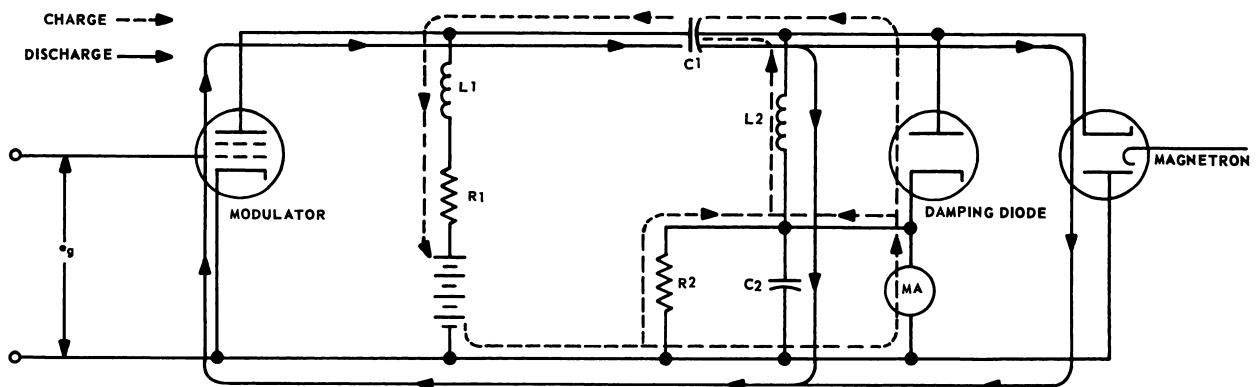


Figure 12-49. Hard-Tube Modulator with Tail Clipping

at the end of the pulse, you need a small inductance. It is evident that the inductance cannot be both large during the pulse and small at the end of the pulse, so a value that is a compromise between the two is used.

**TYPES OF DRIVER STAGES.** The two principal types of driver stages that are used for shaping the pulse applied to the hard-tube modulator are the boot-strap driver and the blocking oscillator.

The boot-strap circuit was widely used in earlier radar sets because it did not require a pulse transformer, a device that was not then perfected. The boot-strap driver circuit forms the pulse by discharging an artificial transmission line through a gas tube. Since the voltages of the pulse required to trigger the hard-tube modulator exceed the voltages that can be present in a gas tube, an amplifier is added between the gas tube and the modulator. To permit "off" tube operation of all three stages, cathode coupling is used. This coupling is unique in that it accomplishes amplification of the pulse without polarity inversion.

In the basic circuit diagram of the boot-strap driver in Figure 12-50, the circuits in heavy lines are those along which pulse formation and amplification progress. One side of the pulse-forming line is connected to  $-400\text{v}$  and the other to  $-1000\text{v}$ . This line charges to  $600\text{v}$  when it is

allowed sufficient time. In this modulator, the pulses are only 500 microseconds apart and the time constant of the charge circuit is about 300 microseconds. The pulse-forming line has time only to charge to approximately 500 volts.

The discharge circuit which the line uses when the gas tube is made to conduct by the trigger pulse is through the parallel combination of the cathode-to-grid resistance of the 829 tube and the  $2.5\text{K}$  cathode resistor in series with the RX233 tube. The resistance of the RX233 is negligible and of the parallel combination is approximately  $Z_o/2$ , where  $Z_o$  is the characteristic impedance of the line. (In some circuits the parallel combination has a value of  $Z_o$ .) With the line discharging through  $Z_o/2$ , one-third of the voltage to which the line is charged appears across the  $2.5\text{K}$  resistor and the grid-to-cathode resistance of the 829 for a period of time determined by the components of the pulse line. This 167-volt pulse applied between grid and cathode of the 829 causes it to conduct heavily since its bias is  $-100$  volts (cathode is at  $-1000\text{v}$  and grid at  $-1100\text{v}$ ).

When the 829 is conducting heavily, the drop across the tube is about 250 volts so that the cathode rises to about  $+300$  volts. This represents a change of about 1300 volts since in the nonconducting period the cathode is at  $-1000$

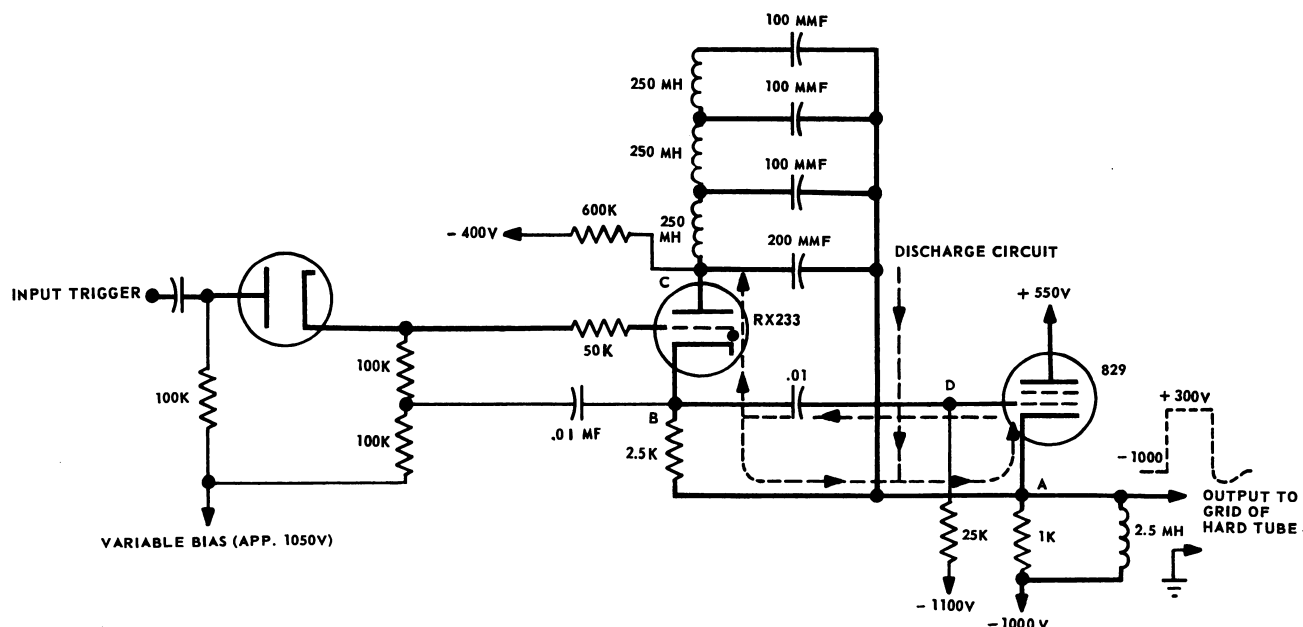


Figure 12-50. The Boot-Strap Driver for Hard-Tube Modulator

volts. The purpose of the 2.5 mh inductance in parallel with the 1K cathode resistor of the 829 tube is to produce the tail at the end of the pulse. The tail results from the collapsing of the magnetic field after the pulse. When the field collapses, current goes through the 1K resistor in the downward direction and causes point A to be more negative than  $-1000\text{v}$  for a brief time. This action insures a rapid cutoff of the hard tube.

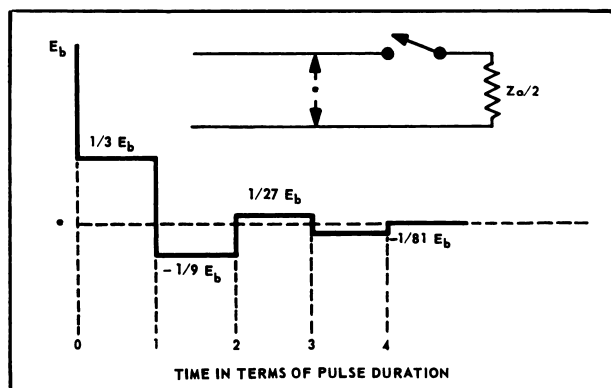


Figure 12-51. Discharge of Pulse Line into  $Z_o/2$

The discharge of a transmission line through an impedance of  $Z_o/2$  is oscillatory, as shown in Figure 12-51. In this diagram you can see that

at the end of the time for a pulse, the voltage across the line swings to  $-E_b/9$ . (In this circuit this equals  $-500/9$  or  $-55$  volts.) Thus the gas tube and the 829 are cut off very quickly. The purpose of the diode in the input circuit is to isolate the trigger circuit from any transients.

A summation of the action at points A, B, C, and D before and during each pulse is as follows:

|        | A               | B               | C                   | D                   |
|--------|-----------------|-----------------|---------------------|---------------------|
| Before | $-1000\text{v}$ | $-1000\text{v}$ | $-500\text{v}$      | $-1100\text{v}$     |
| During | $+300\text{v}$  | $+467\text{v}$  | app. $+467\text{v}$ | app. $+367\text{v}$ |

In a line-controlled blocking-oscillator driver circuit, Figure 12-52, the pulse-forming line controls the duration and the waveshape of each pulse. Normally, the oscillator is cut off by the  $-140\text{v}$  bias. During the interval between pulses the line is charged to  $140$  volts since it is connected to the bias supply through the  $27\text{K}$  resistor, the transformer secondary,  $S_1$ , and  $R_k$  (the cathode resistor of the trigger tube). The trigger pulse, which is about  $150$  volts in amplitude, is sufficient to raise the grid of the blocking oscillator well above cutoff. It is coupled to the grid through the capacitance of the pulse-forming line.

As in an ordinary blocking oscillator, the plate voltage drops when the tube begins to conduct. This drop is coupled back to the grid in opposite phase due to the transformer action and causes the grid to become more positive. This action is

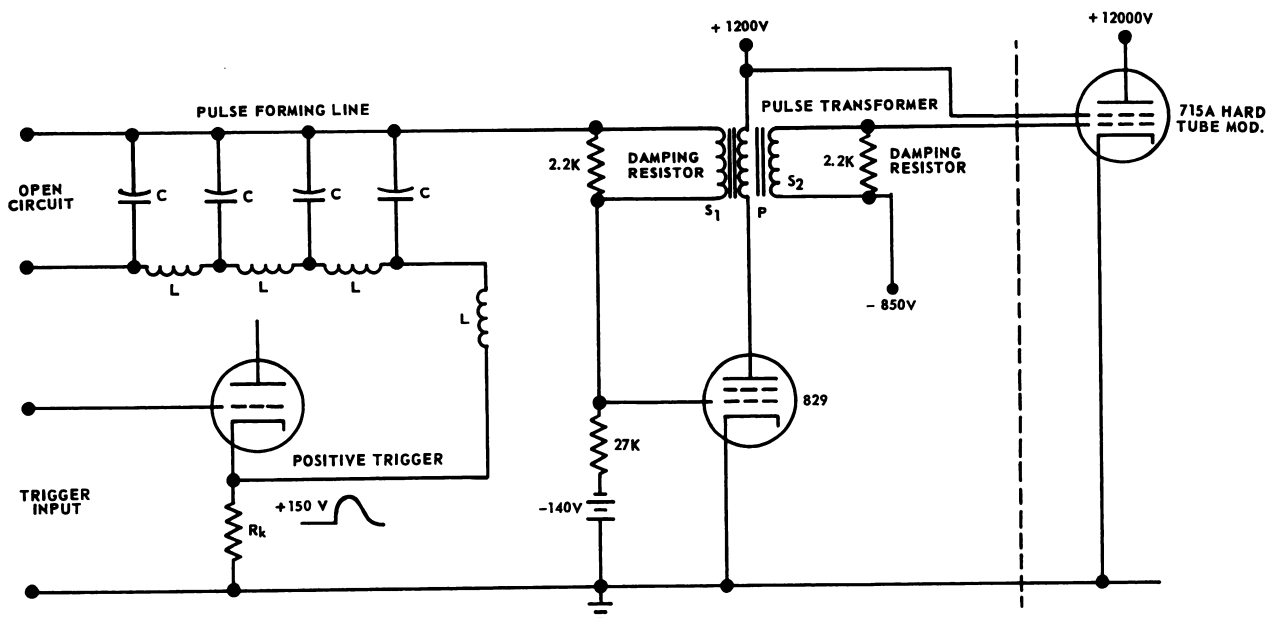


Figure 12-52. Line-Controlled Blocking Oscillator Driver

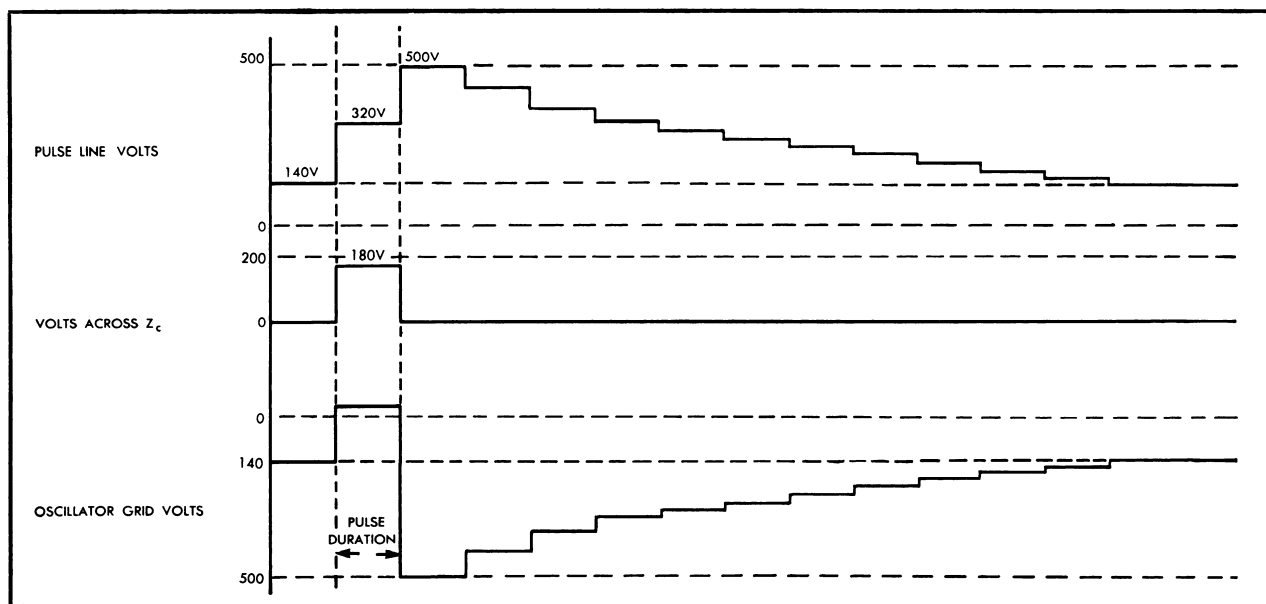


Figure 12-53. Blocking Oscillator Voltages

rapid enough to quickly drive the tube to saturation. In this condition the voltage across the primary P of the transformer is about 1000 volts and that across the secondary  $S_1$  is about 500 volts. Since the pulse line is charged to 140 volts and since it is in series with the secondary voltage of 500 volts, it will charge to the higher voltage through the grid-to-cathode resistance of the oscillator, the secondary  $S_1$ , and  $R_K$ . Further, the impedance of the circuit is matched to the characteristic impedance of the line. Thus, the voltage across the line goes from 140 to 500 volts in two steps as shown in Figure 12-53.

The length of the step at 320 volts is the time it takes the voltage impulse to travel the length of the line, be reflected and return to the sending end. At that time the line is fully charged and current ceases to flow. The current flow through the grid-to-cathode resistance is practically constant during this time and keeps the grid-to-cathode potential steady, insuring little change in the primary voltage. At the time the line becomes charged, current flow ceases through the charging circuit and causes a drop in the oscillator grid voltage. This decreases the plate current and causes a rise in plate voltage. Through the transformer this drop appears as a drop in voltage at the grid and still further reduces conduction.

The action is cumulative, eventually resulting in the tube being cut off quickly. In this con-

dition the line is charged to 500 volts. As it has only 140 volts applied, it starts discharging through the 27K resistor,  $S_1$  and  $R_K$ . As this path has a much higher impedance than the  $Z_0$  of the line, the discharge appears as a series of steps that follows the general exponential curve. The output voltage applied to the grid of the hard-tube modulator is taken from another secondary winding  $S_2$ . This is a 1000-volt pulse. Damping resistors shunt both secondaries.

## TYPICAL MAGNETRON TRANSMITTER

### Block Diagram Analysis

The line-pulsing network in the modulator supplies a negative pulse to the input primary of the pulse transformer, Figure 12-54. The output of the high-voltage power supply determines the amplitude of the pulse, and the characteristics of the pulse-forming network determine its duration. The amplitude will vary generally from a negative 5kv to a negative 15kv. The duration will vary from  $0.5\mu s$  to  $2\mu s$ . The step-up ratio of the pulse transformer increases the amplitude of the signal to a value that is optimum for operation of the magnetron. The impedance of the pulse transformer should match the characteristic impedance of the pulse-forming network to have maximum transfer of power.

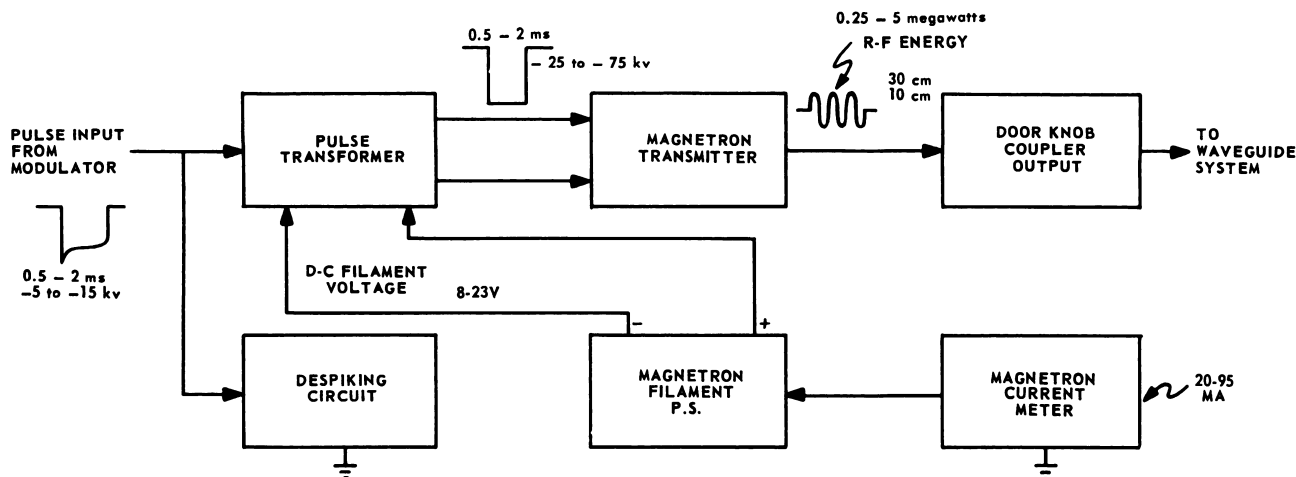


Figure 12-54. Typical Magnetron Transmitter Block Diagram

A high, negative voltage pulse from the secondary of the pulse transformer is applied to the cathode of the magnetron by the bifilar secondary windings. This negative pulse is considerably higher in amplitude but of the same duration as the input pulse. The usual value of the pulse is between a negative 25kv and a negative 75kv, depending upon the type of magnetron used. The conducting impedance of the magnetron is actually reflected back to the pulse transformer and becomes the impedance of the secondary windings. This impedance is determined by the type of magnetron used, voltage applied to the cathode, and strength of the transverse magnetic field.

This secondary impedance directly affects the primary winding impedance as shown by the formula:

$$A = \frac{B}{X^2}$$

where

A = primary winding impedance

B = secondary winding impedance

X = turns ratio, primary to secondary

You can see by using this formula that the impedance match and the power transfer from modulator to transmitter may be affected by several things. Among these are high-voltage power supply output, magnetic field strength, and pulse transformer step-up ratio. Magnetron characteristics and arcing in the pulse-forming network discharging circuit also will affect the impedance

match and power transfer. Although the magnetron transmitter will operate over a relatively large variation of input voltages, it will be most efficient at the voltage for which the circuit was designed. Normally, the magnetrons are, at best, about 50% efficient.

A despiking RC network is generally used in parallel with the primary of the pulse transformer. The despiking network eliminates the undesirable effects of a spike on the leading edge of the pulse applied to the magnetron. The spike is caused by a momentary impedance mismatch before the magnetron starts oscillating. This network is discussed more in detail later in this chapter.

Filament voltage for the magnetron is usually a low DC voltage, although in some sets a 60-cycle voltage is used. The voltage for the filament ranges from 8 to 30 volts and the current from 4 to 10 amperes. The filament power supply is usually a low-voltage, high-current type of power supply. The filament voltage is applied to the magnetron through the bifilar wound secondary of the pulse transformer as illustrated in Figure 12-55.

The impedance of the pulse transformer to the DC or AC filament current is very low. During the time the transmitter is firing, high current is flowing in the secondary of the transformer. The peak current ranges in value from 20 to 50 amperes and lasts only for the duration of the pulse. In order to isolate the filament power supply from the voltage that would be developed across the power supply if the peak current were

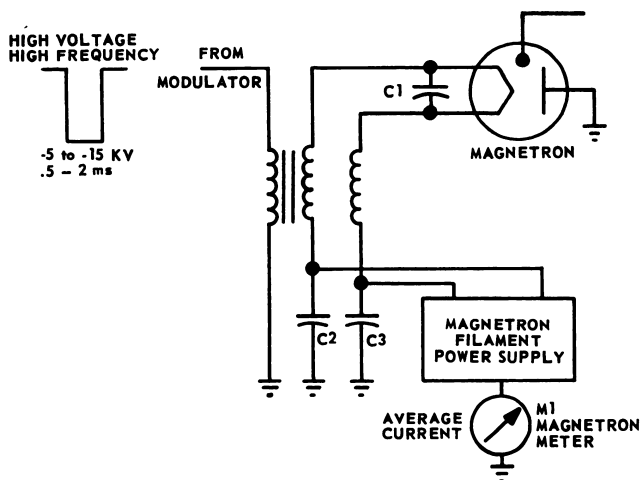


Figure 12-55. Basic Magnetron Filament Circuit

allowed to flow through it, capacitors  $C_2$  and  $C_3$  are placed as shown in Figure 12-55.

These capacitors filter out the current peaks, leaving an average current. The average current is normally 30 to 95 milliamperes. The average current flow from cathode to plate of the magnetron will also flow from ground through the filament power supply and pulse transformer secondary to the magnetron cathode. Now, an ammeter can be placed in series from ground to the filament power supply to measure average magnetron current.

Consider the high-voltage pulse from the modulator as a half-cycle of voltage. Because of the short duration, it is a high-frequency signal. You can see this by the formula:

$$f = \frac{1}{T}$$

where  $T$  is the time for one cycle. Since the negative pulse represents one-half of a cycle, the time for one cycle will be two times the pulse-width. It is possible to isolate the high-voltage pulse from the filament voltage by using high- and low-pass filters. Capacitor  $C_1$ , Figure 12-55, protects the filament of the magnetron. By placing this capacitor in parallel with the filament, both sides of the filament are placed at the same potential for the high-frequency pulse.

The output from the magnetron is coupled from one of the cavities by an inductive loop. A doorknob coupler couples the energy into the waveguide. This type of coupling gives maximum power transfer over a relatively wide range of frequencies.

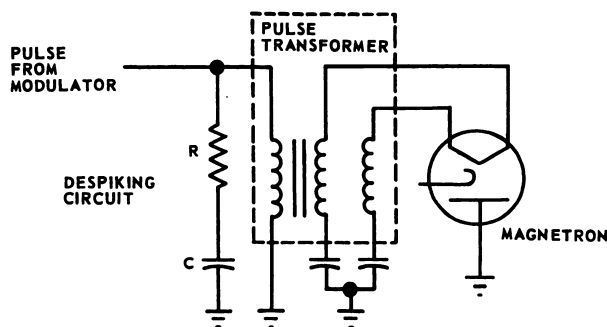


Figure 12-56. Despiking Circuit

### Despiking Circuit

The despiking circuit compensates for the failure of the magnetron to oscillate instantaneously when a negative high-voltage pulse is applied. This circuit may consist of either an RCL or an RC network, as shown in Figure 12-56.

Just before the magnetron oscillates, the impedance of the pulse transformer and magnetron reflected into the ATL discharge circuit is high. The voltage will build up to high values that may cause the magnetron to arc without oscillating. Some precaution must be taken to prevent the magnetron from arcing. The despiking network acts as a low-pass filter in the pulse line, thereby eliminating the high-frequency components of the sharp leading edge of the pulse. Figure 12-57A shows the spike on the leading edge of the pulse. The impedance of the ATL and the impedance of the pulse transformer primary are of approximately the same value when the magnetron is oscillating. Regardless of what the reflected impedance is, the total impedance cannot exceed the value of  $R$  because the total impedance of a parallel circuit cannot exceed that of the smallest resistance.

By selecting a value of  $R$  that is slightly greater than the impedance of the ATL, a fairly good impedance match is obtained for the leading edge of the pulse.  $C$  (Figure 12-56) charges up very rapidly and then the RC circuit has a high impedance. However, by this time the reflected impedance has decreased greatly because the magnetron has started to oscillate. In some ground radar sets a coil is used in series with the ATL to slope the leading edge of the pulse (Figure 12-57B) and therefore ease the magnetron into oscillation.



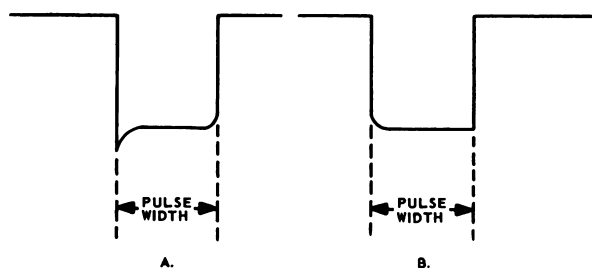


Figure 12-57. Effect of Despiking

## CARE OF MAGNETRONS

Magnetrons are handled with extreme care at the factories where they are produced. Special shipping cartons are provided to insure a maximum amount of protection for the magnetron during shipment. Magnetrons are very sensitive to vibration and shock because the cathode may become off-centered slightly and thus cause the tube to be destroyed when used in a transmitter. Packaged magnetrons (those magnetrons with integral magnets) are damaged when subjected to an external magnetic field or when placed in contact with ferromagnetic materials. Many magnetrons are destroyed by careless handling before they are placed in a transmitter. The following set of rules for handling magnetrons is stressed highly.

1. Use the correct tools for opening shipping crates containing any type of electronic equipment.
2. When it is necessary to store a magnetron for a period of time or to ship it to another destination, be certain that the dust covers are placed over the cathode bushing and r-f output windows. The danger of destroying an expensive magnetron because of accumulated dirt can only be attributed to neglect.
3. Do not strike any surface with the magnetron nor strike the magnetron with any other object during installation, since it may be very harmful to both the magnetron and the magnet. **BE VERY CAREFUL.**
4. Do not try to force the magnetron into position when installing it. This results in shattered ceramic parts and broken seals.
5. Use nonmagnetic wrenches when installing magnetrons. This protects the magnetron from damage, and also lessens the chance of skinned knuckles when working in a magnetic field.

6. Use only the grade of oil specified by the equipment manufacturer for the pulse transformer and magnetron. Failure to do this may result in pitted bushings and eventual failure of the magnetron.

7. Never use steel wool or any other metallic abrasive to clean a magnetron. If cleaning is necessary, use a nonmetallic cleansing agent. Steel particles from steel wool, because of the strong magnetic field, can adhere to insulating parts and thus cause high-voltage breakdown.

By following the preceding simple set of rules, you will prevent many magnetron failures resulting from mishandling.

Some other faults which can cause magnetron failure are:

1. High-voltage standing-wave ratio in waveguide. This causes burned and cracked r-f output windows.
2. Low oil level in pulse transformer. Result of this is heavy internal arcing of magnetron and subsequent damage to cathode and ceramic insulators.
3. Defective ATR and TR tubes, insufficient waveguide pressure, and dirt or moisture in waveguide. These troubles may cause high energy to be reflected back to the transmitter. This would result in a high-voltage standing-wave ratio and a damaged magnetron.

Remember, microwave transmitter tubes are very fragile and subject to damage. Handle these tubes with as much care as you would a CRT.

## VELOCITY-MODULATED TUBES AS LOCAL OSCILLATORS

In radar most receivers use 30- or 60-mc intermediate frequencies. A highly important factor in receiver operation is the stability of the local oscillator which generates the frequency that beats with the incoming signal to produce the i-f. For example, if the local oscillator frequency is 3000 mc, a frequency shift of as much as 0.1% would be a 3-mc frequency shift. This is equal to the bandwidth in most receivers and would cause a considerable loss in gain.

Still another consideration in radar is a receiver that uses a crystal mixer. In such a receiver the power required of the local oscillator is small, being only 20 to 50 mw in the 3000-mc region. Due to the very loose coupling, only about one milliwatt actually reaches the crystal.

A third requirement of a local oscillator is that it must be tunable over a range of several megacycles. This is to compensate for changes in the transmitted frequencies and in its own frequency. In cases where the r-f units are located in a remote position, it is desirable to mount the converter section of the receiver remotely. This insures that the intermediate frequency, rather than the transmitted frequency, is cabled back to the receiver, and makes it necessary that the local oscillator be tuned from a position some distance away, preferably by varying the voltage applied to it.

Because the reflex-velocity-modulated tube meets these three requirements, it is used for local oscillators in microwave radar receivers. The following deals with its operation. In it the cavity resonators are treated as parallel resonant circuits, as was explained in the theory of the resonant cavity in the preceding chapter.

### Theory of Operation

In the circuit of the reflex-velocity-modulated tube, Figure 12-58, note the arrangement of electrodes and the voltages involved for operation. Electrons are emitted by an indirectly heated cathode. These electrons are attracted

by the cavity grids which are more positive than the cathode by the amount of voltage  $E_a$ . The control grid, which is located between the cathode and the cavity grids, has as its purpose the control of electron flow. It also has a *positive* potential which is usually about 200 or 300 volts. The electrons emitted from the cathode travel toward the cavity grids at a velocity determined mainly by  $E_a$ .

Most of the electrons pass through the control grid, the cavity grids, and continue on toward the repeller plate. After passing the cavity grids they come to a region where the electrical field opposes their motion, since the repeller plate is negative with respect to the cathode by the voltage  $E_r$ . This voltage is variable with a maximum of about  $-100$  volts. This, in turn, makes the voltage from repeller to cavity grids 300 or 400 volts and slows down the electrons, causing them to come to a stop. After stopping they reverse direction and pass back through the grids. They are collected either by the control grid, the shell, or the cathode of the tube.

An important consideration is the cause of oscillations in resonant cavities. In most oscillators, oscillations start from some irregularity in current flow, such as a transient that results from voltage being suddenly applied to the tube or from shot effect. With this in mind, assume that the oscillations in the resonant cavities are already taking place. From this assumption examine the source of the energy needed to sustain these oscillations.

With the tank circuit oscillating, a high-frequency voltage,  $e$ , appears between the two cavity grids. This makes the electric field between these grids reverse twice each complete cycle of operation. In this condition as the electrons approach these grids, the electron stream is uniform. The time that is required for the electrons to pass through the short distance between the grids is small compared to the period of oscillations. Electrons that enter the space between the grids when  $e$  is zero will encounter no AC electrical field and will pass on through at the same velocity. The electrons which enter the space when  $e$  makes the left grid negative with respect to the right will encounter a field that tends to accelerate them. The amount they are accelerated is a function of  $e$ . Electrons entering the space when  $e$  is reversed in polarity are decelerated.

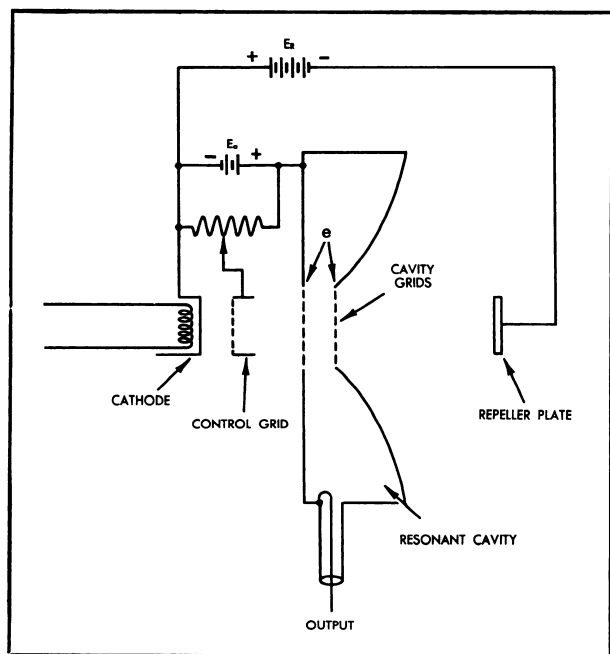


Figure 12-58. Circuit of Reflex-Velocity-Modulated Tube Oscillator

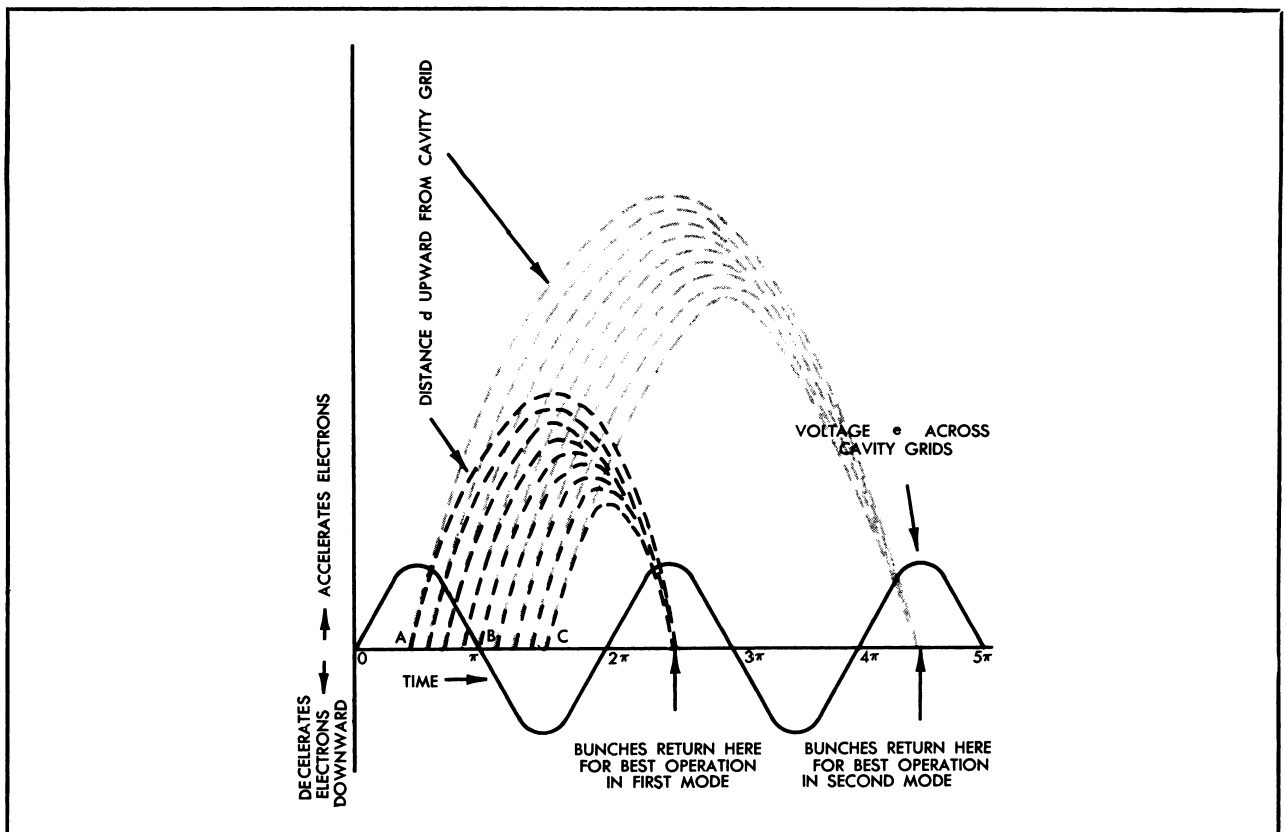


Figure 12-59. Bunching Action of Reflex-Velocity-Modulated Tube

The change in velocity caused by acceleration and deceleration is small in comparison with the original velocity. Those electrons which are accelerated most will travel farther toward the repeller plate before being turned back, while those that are decelerated most will be turned back before approaching very close to the repeller. Thus it is conceivable that with the proper magnitudes of  $e$ ,  $E_a$  and  $E_R$ , electrons returning to the cavity grids will arrive in bunches.

Figure 12-59 shows the position of electrons in the tube at various times during their transit. The zero distance position is midway between the cavity grids. Electron A, which arrives when  $e$  is positive, is accelerated and travels farther before being turned back; electron B is unaffected; electron C is decelerated and turns back after a shorter excursion. Hence, in the diagram these electrons and the ones passing through at intermediate times are shown as arriving back at the grids at the same instant of time. This is the ideal situation, but it is not difficult to see that electrons will return to these grids in a

stream that varies in intensity at the frequency of the oscillations. Therefore, this tube is called a velocity-modulated tube. It is called a *reflex*-velocity-modulated tube because the electrons reverse direction and travel through the inter-electrode space twice.

On the return trip the electric fields set up by the voltage  $e$  again act upon the electrons. Since they are now traveling in the opposite direction, they will be decelerated if they return when  $e$  is positive and accelerated if they arrive when  $e$  is negative. You know that an electron which is accelerated by an electric field has its kinetic energy increased and that this additional energy is taken from the electric field. Also, an electron that is decelerated gives up energy to the electric field. If the bunches of electrons can be made to arrive back at the grid when  $e$  is positive, they will give up energy to the alternating field. For maximum transfer of energy the bunches must arrive when  $e$  is maximum positive.

The question arises as to where this energy originated. You have seen that if the electron

stream from the cathode is uniform, some electrons are accelerated and some are decelerated, on the outbound trip, by the electric field of  $e$ . On the average, as many electrons absorb energy from the field as give up energy to it. Hence, very little net energy is taken from the oscillating circuit during the bunching process. The average kinetic energy of the electron is that imparted to it by the DC voltage  $E_a$ . Thus, you see how energy is taken from the DC electric field and transferred to the AC field to sustain the oscillations.

### Modes of Operation

It is not necessary that the bunches of electrons return to the grids on the first positive swing of  $e$  after they leave them. Figure 12-59 indicates that if the bunches arrive on the second positive swing, the net result is still the same. You can see that the time in transit for the average electron B is three-fourths of a cycle, one and three-fourths cycles, two and three-fourths cycles, etc. In actual practice there are three or four "modes" in which it is possible for the reflex-velocity-modulated tube to oscillate. The black lines in the figure indicate paths of electrons operating in the first mode while the grey lines indicate paths of electrons operating in the second mode. A third mode is possible when the average transit time is two and three-fourths cycles, etc.

The following shows you how the transit time is controlled to produce oscillations in the different modes. If you reason a little, you can understand that since the original velocity of an electron depends on the DC voltage  $E_a$ , and since the distance that the electron travels before turning back and the speed with which it returns depend upon the difference between  $E_a$  and  $E_R$ , it is possible to adjust the two voltages  $E_a$  and  $E_R$  for any of the modes. The voltage  $E_a$  is usually fixed in magnitude since varying it produces greater initial velocity, which in turn causes a farther excursion and a greater return velocity. Since it is not feasible to make  $E_a$  variable,  $E_R$  is variable. For operation in the first mode, the round trip must be completed in the shortest time. This is accomplished by making the repeller plate most negative. For greater time in the interelectrode space, the repeller is made less negative.

Figure 12-60 shows power output and frequency of oscillations as functions of the repeller voltage for three modes of operation. Notice that the frequency at the point of maximum output is the same for all three modes and is the resonant frequency of the cavity. In addition, note that the power outputs for the various modes at the resonant frequency are not the same and that the output is least in the highest mode. This can be explained by examining the factors which limit the amplitude of oscillations and which, in turn, limit the power output.

Power and amplitude limitations are due to overbunching as well as the usual losses in the oscillatory circuit. Overbunching occurs in the following way. As oscillations build up and  $e$  becomes greater, the amount of acceleration and deceleration increases. This causes bunching to occur in a shorter period of time, that is, in a time before the electrons reach the grids on the return trip. This tends to reduce the magnitude of oscillations. In the higher modes of oscillations where the bunches are formed more slowly, the electrons are more susceptible to overbunching. The magnitude of  $e$  that results in overbunching is therefore lower, and oscillations are limited by this action to a lower amplitude than in the lower modes of operation.

As shown in Figure 12-60, the frequency of oscillations in a reflex-velocity-modulated tube is variable to a limited degree in any of the modes of operation by varying the repeller voltage. When the repeller voltage is varied, it causes a bunch to return either a little sooner or a little later than normal. Off resonance, the amplitude of oscillations decreases by an amount depending on the Q of the cavity. In this tube the tuning range is small in comparison with the frequency of oscillations and varies somewhat from one mode to another. It is greatest in the highest mode, because bunching and debunching take place at a slower rate and because greater variation from the ideal time of return is possible without debunching, which would cause the amplitude of oscillations to drop below the usable output level.

Another way to look at it is that in the highest mode the interval between leaving the grids and returning is greater, and the change in period represented by a given change in frequency is a smaller portion of the interval. To illustrate, in the third mode the interval before return must

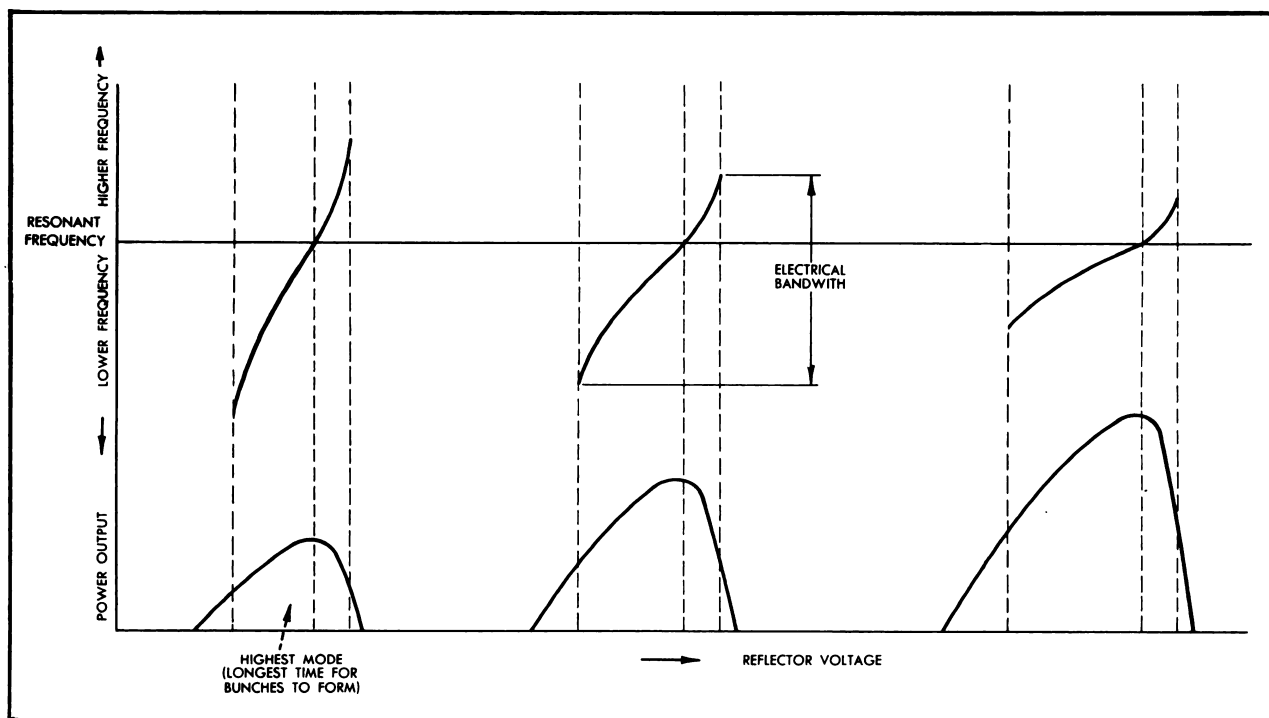


Figure 12-60. Power Output and Frequency Versus Repeller Voltage in Different Voltage Modes for a Reflex-Velocity-Modulated Tube

be about two and three-fourths cycles. A small change in the period of  $e$  would therefore be only  $3/11$  as great a portion of the interval as it would if operation were in the first mode where the ideal time interval is three-fourths of a cycle.

The band of frequencies which can be obtained by varying the repeller voltage lies between the half power points shown in Figure 12-60. This range of frequencies is known as the *electrical bandwidth*. The output curves of the bandwidth are unsymmetrical about the maximum output points. This results from the fact that if  $E_R$  is increased, not only does the bunching voltage  $e$  decrease and cause bunches to form at a later time, but the repeller voltage causes a quicker return. Also, the effects of the two actions add to cause poor bunching at the time the electrons return, resulting in a rapid drop in output on the high side of the hump. At lower voltages, however, even though the bunching voltage  $e$  decreases and causes slower bunching, the decreased repeller voltage causes a later return to the grids. In this way the two effects are counteracting and a greater change in repeller voltage is possible before the output drops below the usable level.

The choice of the point and mode of operation is a compromise among several factors. To begin with, there are three or four modes that have the necessary power output. On the whole, then, it would seem that the correct choice would be the highest mode, for it gives the largest tuning range. The highest mode, however, is too sensitive to a change in voltage to be very well regulated. A change of one volt may cause a change of 0.5 mc in the 3000-mc oscillator. Since the humps are unsymmetrical, the point of operation is usually chosen a little below the point of maximum output. This makes possible the tuning above the operating frequency by a greater amount than if the maximum point were used.

In practice, the reflex-velocity-modulated tube is usually used in conjunction with an automatic frequency control circuit. This circuit controls the repeller voltage in such a way as to maintain the correct intermediate frequency. The details of the discriminator circuit and its operation were discussed in Chapter 9. Keep in mind that the frequency of oscillations is primarily determined by the dimensions of the cavity and that the repeller voltage is effective in making small

| No.  | Name            | Mfr.               | Freq. (Mcps) | Acc. Voltage | Beam Cur. (Ma) | Repeller Voltage | Control Grid Voltage                  | Power Output (mw) | Elec. Tuning (Mcps) |
|------|-----------------|--------------------|--------------|--------------|----------------|------------------|---------------------------------------|-------------------|---------------------|
| K417 | Klystron        | Sperry             | 3000         | 300-600      | 5-30           | +50 to -500      | +5 to +50                             | 150               | 5                   |
| 707A | McNally         | W. E. and Raytheon | 3000         | 250-325      | 25-35          | 0 to -250        | same as acc.                          | 75                | 30                  |
| 726A | Shepherd-Pierce | W. E.              | 3000         | 300          | 22             | -20 to -300      | same as acc.                          | 100               | 20                  |
| 723A | Shepherd-Pierce | W. E.              | 9400         | 300          | 18-25          | -20 to -300      | same as acc.<br>(internal connection) | 20                | 45                  |

Table 12-2. Typical Reflex-Velocity-Modulated Tubes

changes in the frequency. Hence, in most reflex-velocity-modulated tubes there is a coarse frequency adjustment that varies the cavity size in some way. The repeller voltage is the fine frequency adjustment.

### Tubes

Table 12-2 gives some of the operating characteristics of reflex-velocity-modulated tubes. The data in this table will give you an idea of the order of magnitude of tube quantities. There is wide variation between different tubes and different conditions of operation.

The K417 reflex klystron is one of the earlier types that was used for 10-cm operation. One feature of this tube was that in its early application it did not have provision for controlling the frequency through a change in the repeller voltage since both coarse and fine frequency controls changed the cavity grid spacing.

Another 10-cm tube is the 707A (McNally) tube shown in Figure 12-61. In it the cavities are external to the tube and are not evacuated. This makes them susceptible to changes in temperature which results in changes in frequency. To get good frequency stability in it, it is necessary to control the cavity temperature. The coarse frequency control consists of plugs which, when screwed into or out of the cavity, change its size. Fine frequency is controlled by the variable repeller plate voltage control.

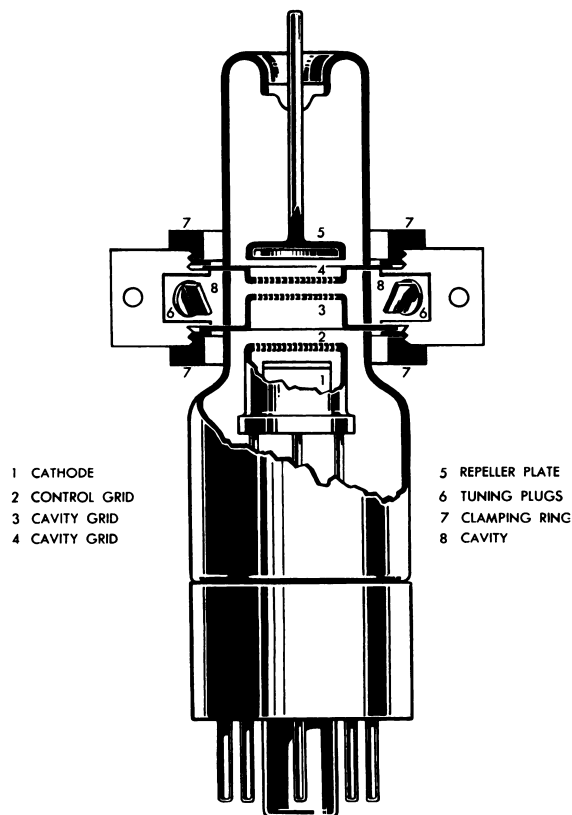


Figure 12-61. 707A (McNally) Tube

The Shepherd-Pierce tube (Figure 12-62) is an all metal tube which is available for both 10-cm and 3-cm operation. In it the cavities are



located inside the tube. Mechanical coarse tuning is accomplished with struts on the side of the tube. The struts are adjusted by a screw which, in turn, varies the size of the cavity. The repeller voltage control also serves as the fine frequency control. The 10-cm and 3-cm type Shepherd-Pierce tubes differ in the shape of the cavity and in the method of coupling the output.

### VELOCITY-MODULATED TUBES AS AMPLIFIERS OF POWER

In the following discussion a brief explanation of the two-cavity klystron, and a detailed explanation of the three-cavity klystron will be given. In addition, this type of power amplifier is compared with the heretofore conventional magnetron. It will be seen that the three-cavity

klystron, like the two-cavity type, possesses a cathode for emission of electrons, an input cavity, a drift tube, an output cavity, and a collector that is positive with respect to the cathode. The three-cavity klystron differs in the addition of a third cavity in the drift tube, in size, and in gain.

### Two-Cavity Klystron

Figure 12-63 shows the construction and essential components of a double-cavity klystron. In this klystron the indirectly heated cathode emits electrons that are focused into a sharp beam by the control grid, which is at a low positive potential. The beam is then accelerated by a very high, positive DC potential on grids  $G_2$  and  $G_3$ , which are very close together and connected to a cavity resonator. Superimposed on this DC potential is

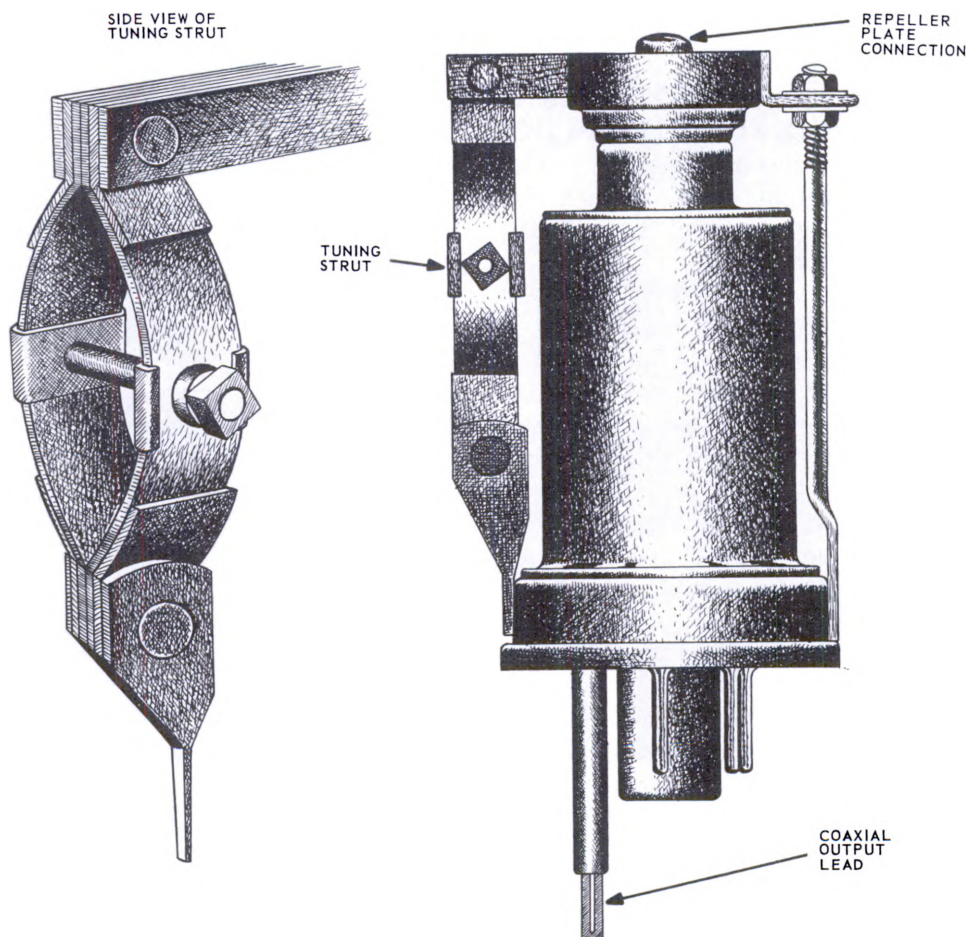


Figure 12-62. External View of Shepherd-Pierce Reflex Klystron

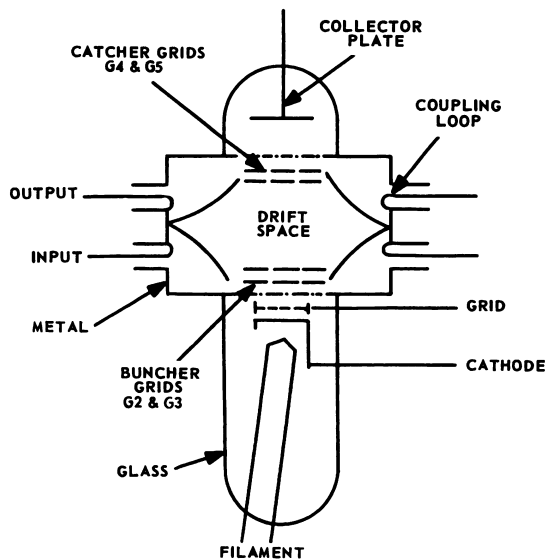


Figure 12-63. Two-Cavity Klystron

an AC voltage that changes the acceleration of the various electrons differently. Those electrons that reach grids  $G_2$  and  $G_3$  at a time when the AC voltage is zero will be accelerated either more or less than normal, depending upon whether the AC voltage is going positive or going negative at the time they reach the grids.

Grids  $G_2$  and  $G_3$  are called *buncher grids* because the electrons, which have been accelerated differently, travel at different velocities through the drift space and will bunch at some point. If these bunches of electrons arrive at grid 4 and grid 5 and are there decelerated, they deliver energy to the cavity resonator connected to these grids. Since the function of grids 4 and 5 is to absorb, or catch, the energy contained in these bunches, they are called *catcher grids*. Electrons that continue through the catcher grid hit the collector plate and are returned to the cathode through an external circuit.

The bunching action that occurs in the klystron is illustrated by the Applegate diagram in Figure 12-64. In it the lines all start at the top of the diagram at the buncher grids; they represent the paths of the various electrons in their travel through the drift space. The AC voltage waveshape at the top represents the AC component of the voltage on the buncher grids; the horizontal axis represents time. The lines are therefore tilted to the right in their passage from

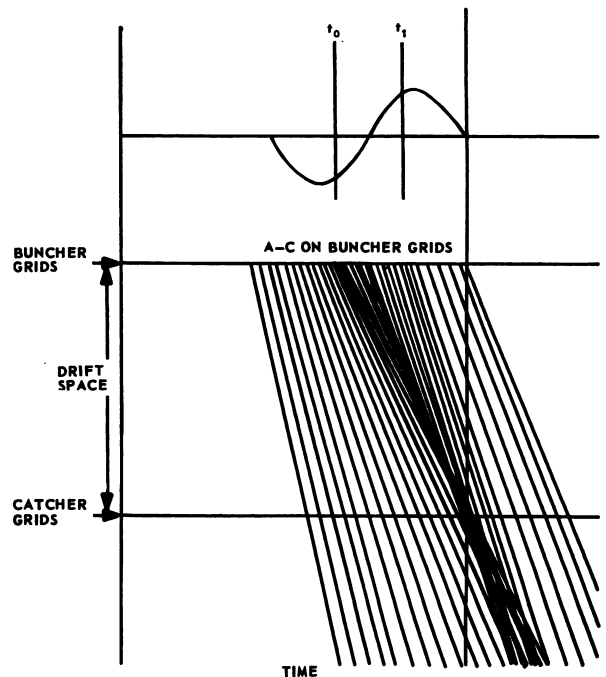


Figure 12-64. Two-Cavity Bunching Action, Applegate Diagram

buncher to catcher grids. The difference in the slope of the various lines as they leave the buncher grids represents the different velocities of the electrons, which are due to the different accelerations imparted to these electrons by the AC component of the voltage on the buncher grids. Note that the electrons which arrive at the buncher grids between times  $t_0$  and  $t_1$  form a bunch at the catcher grids. There is only one bunch of electrons in each cycle; hence, the frequency of bunches arriving at the catcher grids is the same as the input frequency.

For the klystron to function most efficiently, the drift space should not be too long. If the distance between catcher and buncher grids is made too great, the bunches disappear or tend to merge, thus making the beam no longer usable. There is a certain spacing that gives best results with given DC and AC voltages. Since the spacing (drift space) is fixed in a tube, the voltages must be accurately adjusted for the tube to function properly. Also, the resonant circuit of the catchers must be tuned to the correct frequency. The strength and phase of the oscillation must be such as to absorb maximum energy from the electron bunches.

The AC, which is superimposed upon the DC applied to the buncher grids, may be applied through the input terminal and then by a coupling loop to the buncher cavity. The output may be taken from the catcher with a coupling loop through the output terminal. The klystron power amplifier may be used in pulse operation by applying in pulses the DC voltage, the r-f voltage, or both.

### Three-Cavity Klystron

In this type the cathode is heated by an AC operated heater. The heater and cathode assemblies, with leads brought out at the base of the tube, are insulated from the entrance to the drift tube by a glass insulating sleeve. The input cavity is located close to the drift tube entrance. The middle cavity is located farther along the drift tube, and the output cavity is at the other end of the drift tube. The collector is beyond the output cavity. The entire drift tube assembly, the three cavities, and the collector are at ground potential. This eliminates hazards and complications involved in tuning the cavities, and in liquid cooling of the body and the collector. With the drift tube assembly at ground potential, the cathode is pulsed with a negative voltage to accelerate electrons from the cathode toward the drift tube entrance. The negative pulse is supplied from a pulse-forming network and is stepped up to the desired value by a pulse transformer.

In high-power klystrons the amplitude of the cathode pulse is measured in hundreds of kilovolts, and the peak current flowing for the duration of the pulse is measured in hundreds of amperes. The high values of accelerating voltages produce dangerous amounts of X-ray radiation, particularly around the collector, where most of the electrons strike. To protect maintenance personnel from harmful radiation effects, the entire klystron is housed in a heavy lead shield. Because of the great current density and the distance of electron travel, electromagnetic focusing is used to control the electron stream.

The choice of cathode requires much consideration. For a desired r-f power output, a certain amount of DC power must be available. If a peak r-f output of one megawatt is desired, with an expected efficiency of 33%, a peak DC power input of three megawatts is required. The DC input power is the product of beam voltage and

current. The same amount of power can be supplied at high voltage and low current or at low voltage and high current. There is a practical limit to the maximum pulse voltages that can be generated consistently at the PRF's common in modern radar systems. On the other hand, an electron beam of very great density is difficult to control and produces severe focusing problems. The beam voltage and current values used are therefore a compromise between high beam voltage and current density.

The relation between beam voltage and beam current determines the characteristics of the cathode. The cathode characteristic, which is of chief importance, is *perveance*. This is the ratio of cathode current to the beam voltage raised to the three-halves power. It is a measure of the ability of the cathode to emit electrons at various beam voltages, and is determined by the emitting material of the cathode, the cathode area, and the operating temperature. The impedance presented to the pulse transformer secondary by the klystron during conduction is equal to the ratio of beam voltage to beam current. Since this ratio is determined by the perveance of the cathode, the perveance determines the impedance reflected through the pulse transformer to the pulse-forming network supplying the cathode pulses.

Current flows in the pulsed klystron for the duration of the cathode pulse. Between pulses no potential difference exists between the cathode and the drift tube entrance and, therefore, no current flows. The remaining discussion will apply only to the period of conduction, or during the cathode pulse time. When the cathode is pulsed negatively, electrons are accelerated toward the drift tube entrance. The total acceleration given to the electrons by the cathode-to-drift-tube potential, resulting from the negative cathode pulse, occurs while the electrons are in transit between the cathode and the drift tube. Figure 12-65A presents an approximate picture of the electric field lines (shown dotted) existing between the cathode and the drift tube entrance.

The electric field lines do not extend into the drift tube for any appreciable distance. With only the electric field present, the electrons would tend to follow the field lines. Under this condition, the electrons would travel from the cathode to the drift tube entrance, and the energy imparted to the electrons by the cathode pulse

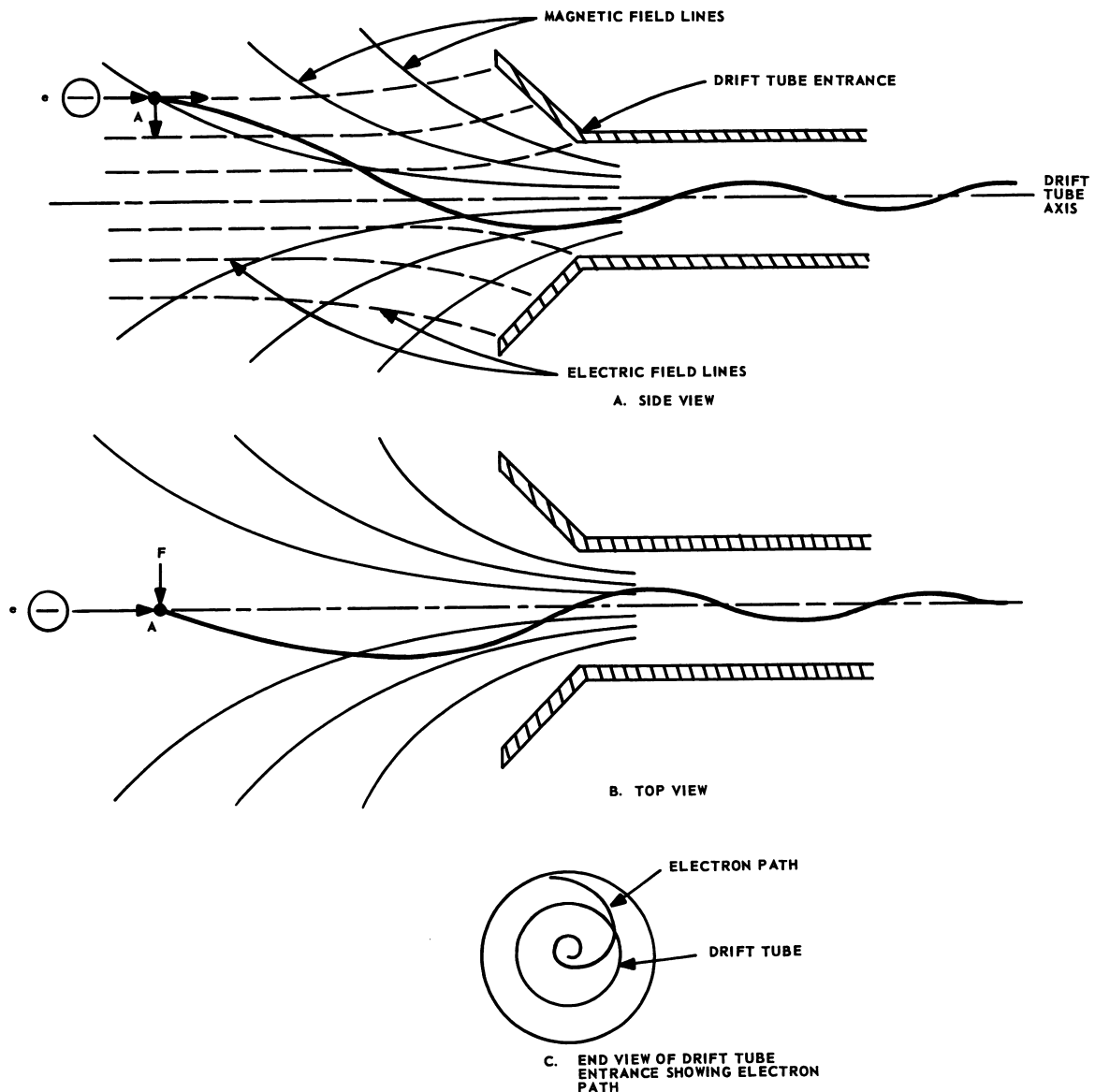


Figure 12-65. Convergence of Electron Stream at Entrance of Drift Tube in Klystron

would be dissipated at the drift tube entrance as heat generated by electron bombardment of the metal surface. To avoid this, a magnetic focusing system is employed to divert the electron stream into the drift tube. Such a field must be symmetrical about the drift tube axis. A few lines of such a field (solid lines) are shown superimposed on the electric field lines of Figure 12-65A. The focusing action at the drift tube entrance can be seen by referring to Figure 12-65B. This figure shows a cross section of a magnetic field, which is symmetrical about the drift tube axis.

The path of a single electron under the influence of the electric field alone is shown as a heavy solid line.

At point A in Figure 12-65B, where the electron path intersects a magnetic field line, the line is resolved into its components. One component (axial component) is seen to be parallel to the direction of the electron path and the other (radial component) is at right angles to the electron path. The radial component force causes the electron to rotate out of the plane of the paper, around the axis of the drift tube. This

tangential force causes the electron path to become helical as it approaches the drift tube entrance. At the drift tube entrance the magnetic field intensity is much greater than that at the cathode, causing the spiral of the electron path to become tighter about the drift tube axis. Also, at the drift tube entrance, the magnetic field lines are almost entirely axial.

The tangential component of the electron velocity in the presence of the axial field causes the electron to be accelerated radially inward. The electron path as viewed from the cathode end of the drift tube is shown in Figure 12-65C. The greater the initial distance (perpendicular) of the individual electrons from the drift tube axis, the greater the forces tending to direct the electrons toward the drift tube axis. Electrons released from the cathode at or near the drift tube axis experience relatively slight effects of the magnetic focusing field and travel an almost straight line path from the cathode to the drift tube entrance. The over-all effect of the magnetic field is to force electrons emitted from the cathode into the drift tube, thus keeping to a minimum the number striking the metal surrounding the drift tube entrance.

Once the electrons enter the drift tube, they are no longer under the accelerating effects of the cathode-to-drift-tube potential, and they coast, or drift, at constant speed until they encounter the fields across the gaps of the three resonant cavities arranged along the length of the drift tube. The axial field produced by the entire focus coil assembly extends the length of the drift tube. The electrons within the drift tube travel parallel to the axial magnetic field. Whenever, for any reason, an electron is deflected from the axial path, a radial component of velocity is introduced that causes the electron to spiral about the axial magnetic lines. It is impossible to provide the degree of focusing required to prevent all electrons from striking the walls of the drift tube. By proper adjustment of the magnetic field, however, it is possible to hold the number to a practical minimum.

Current produced by electrons striking the drift tube walls is referred to as body current. A high body current is undesirable since it indicates that a large number of electrons are giving up energy as heat in collisions along the drift tube walls rather than contributing energy to the r-f output cavity. Several factors influence

the amount of body current produced. The most important is the focusing adjustments. Another is the interaction of electrons in a dense electron stream. Since all electrons are negatively charged, they tend to repel each other; this produces beam spreading and subsequent collisions along the drift tube walls. Another effect is also produced by the concentration of negative charge within an electron stream.

Electrons along the outer edges of the stream see a concentration of negative charge at the center of the stream (corresponding to the drift tube axis). The drift tube walls, being at ground potential, appear positive with respect to the concentrated charge along the axis of the tube. The fringe electrons are, therefore, accelerated toward the drift tube walls because of this apparent potential difference between the center of the drift tube and the walls.

As previously stated, electron behavior is influenced by the cavity gap voltages once the electrons have entered the drift tube. Cavities of the gridless type are used because of the unusually great electron velocities produced by the high beam voltage. The edges of the cavity gaps are flush with the drift tube wall to avoid intercepting electrons in the beam.

Cavity resonator construction for klystron tubes is of two primary types. One type is built into the tube and is evacuated as part of the tube structure. Tuning of this type of cavity is performed by some type of bellows-driving mechanism that alters the physical size of the cavity. Another type of cavity is mounted externally to the evacuated tube section. The portion of the drift tube comprising the resonant cavity gap is a cylinder of ceramic or other low loss insulating material. The tuned cavity can be removed from the drift tube in such an assembly.

The chief factor in cavity gap spacing is electron transit time. The ideal condition is zero transit time but, of course, this is not possible at the desired operating frequencies. There is a finite time interval involved in the passage of electrons across the gap of the resonator. The time required for passage is determined by the initial electron velocity upon entering the gap and upon the width of the gap. The operating frequency determines the passage time required in terms of a complete r-f cycle. If the width of the gap is made very small to reduce transit time, the cavity impedance is lowered, since this is



equivalent to increasing the capacitance of a tuned parallel-resonant circuit. Also, if the gap is made very small, possibility of a breakdown across the gap exists, because of the large r-f voltages developed.

In addition, with a small gap there is heavy "loading" of the cavity. This is a result of secondary electrons, emitted around the cavity gap, being driven across the gap by the r-f potentials across the gap. Movement of these electrons under the influence of the r-f field across the gap is an energy drain on the cavity and performs no useful function. This waste of energy is reflected as a reduced  $Q$  of the cavity with associated losses in power. A cavity gap that produces a transit time of a quarter-cycle or slightly more at the operating frequency provides satisfactory beam coupling and, at the same time, avoids the undesirable effects of too small a gap spacing.

The drift tube diameter is also a compromise. While a small diameter gives a high degree of coupling between the cavities and the electron beam, it complicates the focusing problem. A large tube diameter simplifies the focusing problem but reduces beam coupling. However, the diameter of the drift tube must be less than the cutoff dimension at the operating frequency. This is necessary to avoid standing waves along the drift tube. The drift tube diameter is, therefore, a compromise value that gives satisfactory beam coupling to the cavity gaps and at the same time permits practical focusing of the beam.

The output of any klystron (regardless of the number of cavities employed) is developed by velocity modulation of an electron beam. Velocity modulation, in turn, produces electron bunching. The bunching process is a result of the velocity modulation, and is a function of the time required for electrons to travel the distance from the input cavity to the output cavity. The electrons that are initially accelerated by the beam voltage (considered to be DC for the duration of the cathode pulse) are acted upon by r-f fields developed across the input and middle klystron cavities.

Some electrons are accelerated, some are decelerated, while others are unaffected, depending on the amplitude and polarity of the cavity r-f voltages when the electrons cross the cavity gaps. For the interval during which the electrons travel from one cavity to the next, the accelerated

electrons tend to overtake the decelerated ones, thus causing the formation of groups, or bunches, of electrons. As a result, bunches of electrons arrive at the output cavity at the proper instant of each r-f cycle to deliver energy to the output cavity. Transit time is thus seen to be a necessity rather than a hindrance to the klystron operation.

The bunching process is best explained by reference to Figure 12-66, an Applegate diagram of the bunching process in a three-cavity klystron. The vertical axis represents the distance measured along the drift tube. The horizontal axis represents time measured either in millimicroseconds or in degrees of an r-f cycle. The diagonal lines are electron paths plotted as a function of time and of distance along the tube.

The slope of a line is proportional to the velocity of the electron whose path is described by that particular line. The steeper the slope, the greater the electron velocity. A change in slope, therefore, indicates a change in velocity. An increased slope indicates an increase in velocity, and a decreased slope indicates a decrease in velocity. Between the cathode and the first cavity the electrons emitted from the cathode receive equal accelerations and therefore travel at equal velocities. All electrons entering the input cavity gap are, therefore, seen to have equal slopes.

A sinusoidally varying r-f field exists across the first (input cavity) gap as a result of the signal input supplied from an external r-f source. The input cavity is resonated at the input frequency. The effects of the r-f voltage across the input cavity gap on individual electrons depend upon the instantaneous magnitude and polarity of the r-f gap voltage. The word "instantaneous" implies that the transit time is zero. As previously explained, this is not true, since an appreciable portion of an r-f cycle (over one quarter) elapses during the electron transit across the gap. More correctly, the average magnitude and polarity of the r-f cycle during the electron transit must be considered.

An electron we will designate as electron 1 in Figure 12-66 crosses the input cavity gap when the potential is zero. The velocity of electron 1 is unaffected by the input gap voltage and there is no change in its velocity. This is shown by no change in the slope of the line describing the path of electron 1. Electron 3 crosses the input



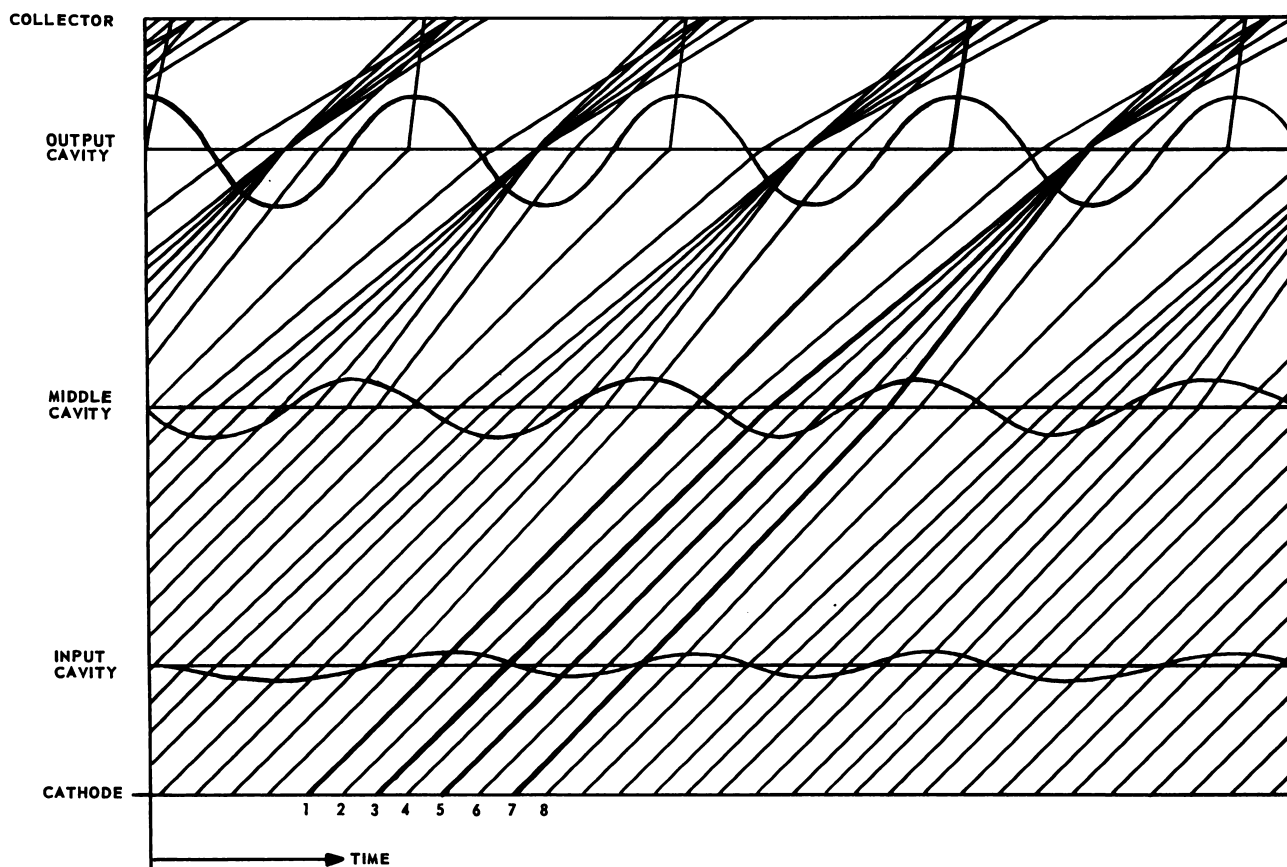


Figure 12-66. Klystron Bunching Graph

cavity gap when the r-f voltage is maximum in a negative direction. This electron is decelerated during its passage through the gap and emerges from the gap traveling at a reduced velocity. This is indicated by a decrease in the slope of line 3 beyond the input cavity.

Electron 5, like electron 1, passes through the gap when the r-f signal is zero and therefore emerges with no change in velocity. Electron 7 crosses the input gap when the r-f voltage is maximum in a positive direction. This electron is accelerated during its transit through the input gap and emerges from the gap at increased velocity. This is indicated by an increase in the slope of line 7 beyond the input cavity gap. Electrons 2 and 4 cross the input gap when the r-f voltage is negative and are decelerated by an amount proportional to the amplitude of the r-f voltage. Electrons 6 and 8 cross the input gap when the voltage is positive and are accelerated by an amount proportional to the amplitude of the r-f voltage.

When the electrons pass the input cavity gap, they enter the drift tube section between the first and second cavities. In this region there are no DC or r-f fields and the electrons travel at whatever velocities they acquired as a result of the combined effects of the beam voltage (cathode pulse) and the r-f voltage across the input cavity gap. In passing through the first gap, electrons emitted from the cathode at successive intervals are accelerated or decelerated as explained above. This process is called *velocity modulation*. In the drift tube section, between cavities one and two, the velocity modulation produces *density modulation* of the beam.

Electron 3 leaves the cathode earlier than electron 5. Initially, both have the same velocity. Since electron 3 is decelerated in passing through the input cavity gap and electron 5 is unaffected, electron 5 will overtake electron 3 if the two drift for a time in a space where no fields are present. Electron 7 is emitted from the cathode at a later time interval than electron 5, but both

have the same initial velocity. Electron 5 is unaffected by the r-f voltage across the input cavity gap, but electron 7 is accelerated. Therefore, electron 7 will eventually overtake electron 5 if they travel undisturbed for a sufficient time. The over-all effect is that electrons crossing the input cavity gap before and after electron 5 tend to converge on, or group around, electron 5. This is slightly evident at the middle cavity.

Velocity modulation of the DC beam by the input cavity does not produce serious loading of the input cavity. The fact that a DC beam is being velocity-modulated means that the number of electrons crossing the input cavity gap when the r-f voltage across the gap is negative is equal to the number crossing the gap when the r-f voltage across the gap is positive. If electrons cross the cavity gap when the r-f voltage across it is negative, energy is supplied to the cavity by the electrons, because in slowing down the electrons give up kinetic energy to the gap. If the electrons cross the cavity gap when the r-f voltage across it is positive, energy is supplied to the electrons from the cavity, since the electrons are accelerated. The r-f input to the first cavity, therefore, is of sufficient magnitude to overcome the inherent losses in the cavity in maintaining the desired r-f voltage across the cavity gap.

Only a small degree of density modulation (bunching) occurs within the electron beam in the interval of travel from the input cavity to the middle cavity. The amount is very small compared to the degree of bunching required at the output cavity. The amount of density modulation is sufficient, however, to excite the middle cavity and, because of the high  $Q$  of this cavity, to maintain a large oscillatory voltage across it. The voltage thus produced across the middle cavity gap is much larger than that across the input gap. This voltage is responsible for most of the velocity modulation, and subsequent density modulation, produced within the klystron.

There is a negligible net transfer of energy from the beam to the middle cavity, or from the cavity to the beam. The amount of energy coupled to the cavity from the beam is the small amount necessary to overcome the cavity losses in maintaining the oscillatory r-f voltage across the gap. The degree of density modulation present in the electron stream at the middle cavity, as a result of the velocity modulation at the input cavity, is so slight compared to that required

at the output cavity that it can be considered as DC for practical purposes. Therefore, the net energy delivered to, or taken from, the electron beam is quite small, as explained for the input cavity.

The large voltage across the middle cavity gap produces a much larger degree of velocity modulation than that of the input cavity. This is evident in the Applegate diagram of Figure 12-66 by sharper changes in the slopes of the electron paths crossing the middle cavity. Electrons 1 and 5 again cross the gap when the r-f voltage is zero, and continue with no change in velocity. Electrons 2, 3, and 4 cross the gap when the r-f voltage is negative and are slowed down, as shown in a decrease in slope of lines 2, 3, and 4. Electrons 6, 7, and 8 cross the gap when the r-f voltage is positive and are accelerated, as shown by increases in the slopes.

After the electrons cross the middle cavity gap, they once more drift in the field region between the middle cavity and the output cavity. The bunching process continues in this region, but at a more rapid rate than that between the input cavity and the middle cavity because of the greater degree of velocity modulation imparted to the beam by the middle cavity. The effect of velocity modulation is shown at the output cavity. Electrons 3, 4, 5, 6, and 7 are bunched and cross the output cavity gap when the gap voltage is maximum negative. Under this condition, maximum energy transfer from the electrons to the output cavity occurs.

The energy given up by the electrons is the kinetic energy of high velocity, imparted to the electrons by the beam voltage (cathode pulse). Giving up this energy produces a decrease in velocity as shown by the sharp changes in slopes at the output cavity. Electrons 2 and 8 did not receive sufficient velocity modulation to cause them to bunch with the others. Electron 1, having traveled unaffected through the input and middle cavities, reaches the output cavity when the r-f voltage across it is maximum positive. The voltage imparts to the electron a sharp acceleration. This is shown by the sharp increase in slope. This acceleration of electron 1 represents a load on the cavity, since energy from the cavity is required to accelerate the electron.

Optimum bunching reduces the number of such electrons to a minimum. The bunching process

pictured in Figure 12-66 is ideal, particularly at the output cavity as the bunched electrons are shown arriving at the output cavity at the same instant. This is not true in practice. The passage of an electron bunch at the output cavity occurs over a period covering a considerable portion of a half-cycle. Also, the history of only eight electrons is discussed. Actually, billions of electrons are emitted from the cathode during the time interval between the emission of electrons 1 and 8. The effects of mutual repulsion produced by such a dense electron stream make it impossible to approach, in practice, the degree of bunching shown in Figure 12-66.

Since it has been stated that the bunching process is a function of time, it would appear that if an electron beam is velocity-modulated by a small r-f voltage and the electrons are permitted to drift for a sufficiently long time, the desired degree of bunching could be obtained with a minimum of input power. The defocusing effects previously discussed prevent this, however. The debunching effects of a dense electron stream would spread the beam in the drift space after a short time interval. The drift time is therefore limited to a time interval corresponding to only a few r-f cycles, and input power is adjusted to produce the desired bunching in this time interval.

The three-cavity klystron is the equivalent of a pair of two-cavity klystrons in cascade, and is referred to as a cascaded amplifier klystron. The three-cavity klystron is more efficient than a pair of two-cavity klystrons in cascade, since the middle cavity replaces the output cavity of one pair and the input cavity of the other. In this way cavity losses are cut in half and the further division of power produced is avoided. (A maximum of 50% efficiency is realized when a matched source and load are coupled.) Another advantage is the simplified circuitry involved. There are only three tuned circuits instead of four; only one electron beam is used, and power supply provisions are also simplified.

**SUMMARY.** The mechanics of bunching within a klystron cause the output power to be critically dependent on input power. For a given frequency, drift tube length, and beam voltage, there is only one value of input power that gives maximum output if all three cavities are tuned for maximum gain. The degree of bunching achieved in a given drift tube space depends on

the amount of velocity modulation imparted to the beam electrons while crossing the gaps of the input and middle cavities. The velocity modulation depends on the amount of r-f voltage appearing across the input and middle cavity gaps.

The greater the voltage developed across the input cavity gap, the greater the velocity modulation produced at this gap and the greater the degree of density modulation produced in the interval of electron travel from the input cavity to the middle cavity. The excitation of the middle cavity depends on the amount of density modulation in the beam arriving from the first cavity. The greater the density modulation seen at the middle cavity gap, therefore, the greater the r-f voltage developed across the middle cavity gap, and the greater the velocity modulation imparted to the beam. If the input power is at the optimum value, the maximum bunching point will occur at the output cavity gap.

If the input power is too low, there is insufficient velocity modulation of the beam, and maximum bunching will occur slightly later, or at a point beyond the output cavity gap. If there is too much input power, too large a degree of velocity modulation is produced, and maximum bunching occurs before the velocity-modulated electrons reach the output cavity. The amount of bunching at the output cavity determines the power delivered to the cavity from the electron beam. Either of these conditions, overdriving or underdriving the input cavity, produces maximum bunching in the klystron at some point along the drift tube other than the output cavity gap with a resulting decrease in output power. A curve of klystron power output versus power input is shown in Figure 12-67.

**PERFORMANCE.** Another consideration is expressed in the curves of Figure 12-67. The power output of a three-cavity klystron can be increased beyond that corresponding to maximum gain (all three cavities tuned to the same frequency). The bunches of electrons arriving at the output cavity contain components contributed by velocity modulation produced both at the input cavity and at the middle cavity. With all three cavities tuned to the same frequency, the bunched current contributed by velocity modulation at the input gap is not ideally phased with respect to the velocity modulation contributed by the middle cavity; therefore, the bunching components

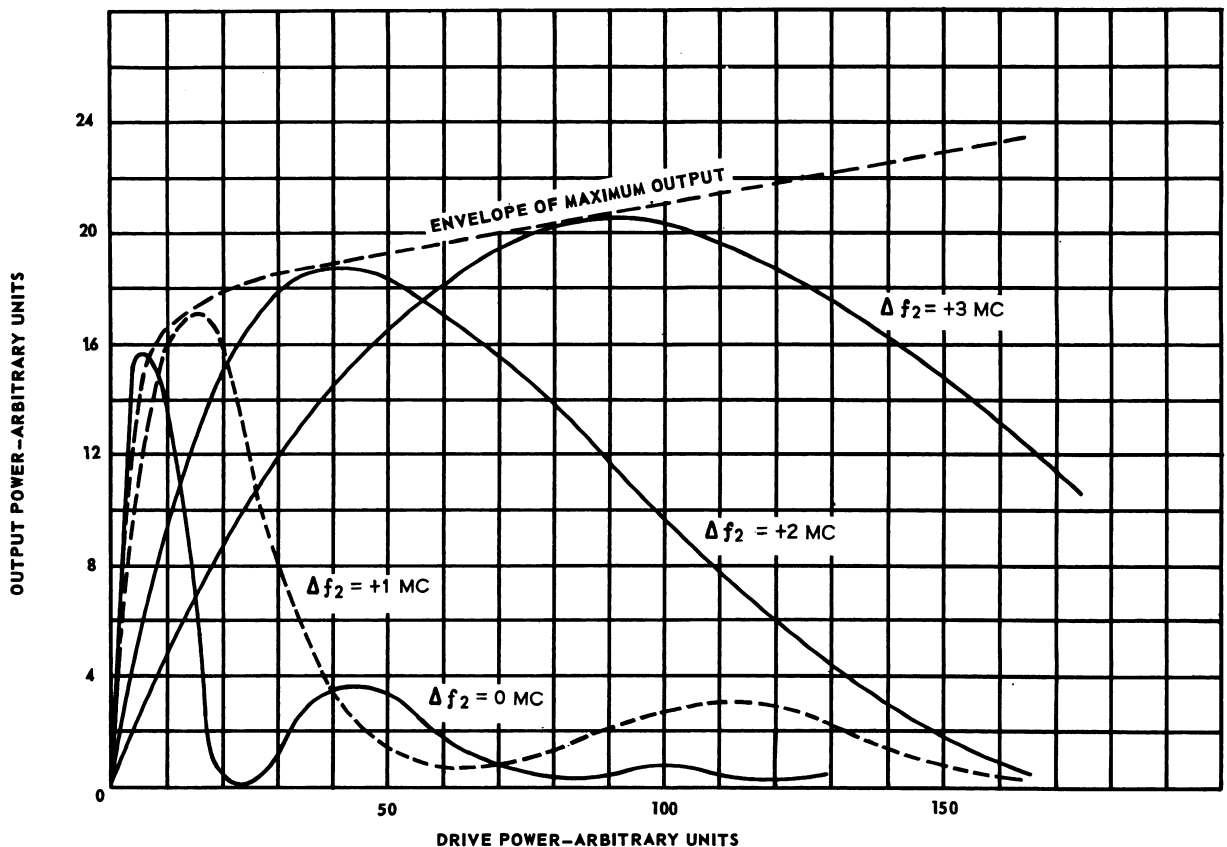


Figure 12-67. Graph of Klystron Input Versus Output Power

of the input cavity do not add with those of the middle cavity to produce optimum bunching at the output cavity.

By detuning the middle cavity the components contributed by the input cavity are phased more advantageously with respect to those contributed by the middle cavity, and the bunching at the output cavity is increased. When the second cavity is detuned to change the phase of the r-f voltage across it with respect to that across the input cavity, the amount of density modulation required to maintain a given r-f voltage across the middle cavity gap is increased. Since the density modulation appearing at the middle cavity is a result of the velocity modulation contributed by the input gap, the power supplied to the input gap must be increased to supply the required increase in velocity modulation. Further detuning of the second cavity results in a further increase in output power, provided the input power is increased to the new required value.

The curves of Figure 12-67 show the results obtained with various amounts of detuning of the middle cavity. Notice that the increase in output power is not proportional to the increase in input power required to maximize the output for a particular degree of middle cavity detuning. Increases in output power obtained by detuning the middle cavity and increasing the input power are obtained at a sacrifice of gain. At the higher power outputs obtained by detuning the middle cavity, the output curve is much broader, and the klystron output is less critically dependent on variations of input power.

The final major consideration in power klystron performance is the shape of the cathode voltage pulse. The amplitude of the pulse determines the transit time of the electrons from the cathode to the collector. Any change in pulse amplitude affects the electron transit time. This, in turn, affects the points of occurrence of maximum electron bunching in the region of the output cavity. If the r-f input to the klystron is

constant and the cathode pulse amplitude varies over the duration of the pulse, the points of occurrence of maximum bunching will vary with pulse amplitude. This results in phase shift of the r-f signal output with respect to the input during the cathode pulse.

Since the transit time from the input cavity to the output cavity is usually equal to the time of several r-f cycles, a small percentage phase shift may still be undesirably large. Any phase shift is undesirable in a moving target indicator system, which yields information based on pulse-to-pulse phase comparisons. Phase shift of the r-f output must therefore be held to a value that will not limit the moving target resolution possible in the cancellation system employed in the equipment. Thus the cathode voltage pulse must be flat-topped.

### Klystron Versus Magnetron

The advantages of the klystron as compared to the magnetron are many. First of all, physical considerations are simplified. In the magnetron the cathode and the anodes are included in the frequency determining system and are a compromise between desired power output and required size. In the klystron, the cathode and collector are external to the frequency determining fields and can be designed independently of the r-f section to handle the desired power. Cooling methods are simplified, since most of the heat generation occurs in the collector and can be dissipated by a cooling system without consideration of the r-f portion of the tube. Figure 12-68 illustrates a three-cavity klystron.

From the standpoint of electrical performance, the klystron again excels. The magnetron is a power oscillator and is the frequency determining element of the radar system. Its frequency stability is dependent on the combination of a magnetic and an electric field, either of which is subject to variations over short periods of time. In a moving target indicator (MTI) system, discussed in Chapter 15, the stable local oscillator is continuously varied by an AFC system to track with the magnetron. A lock (transmitter) pulse is supplied to the coherent oscillator at the beginning of each output pulse to provide a reference phase for the echo returns.

The pulse-to-pulse frequency of the magnetron is subject to considerable variation for which the AFC system cannot compensate instantaneously.

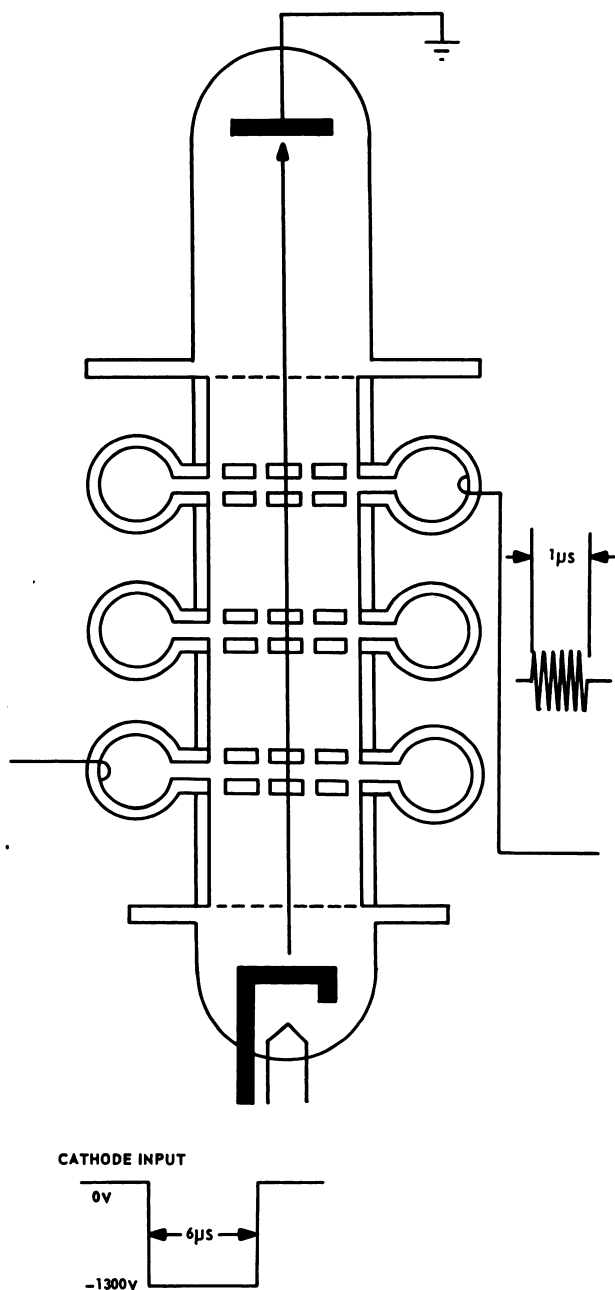


Figure 12-68. Three-Cavity Klystron

As a result, pulse-to-pulse phase differences occur because of magnetron frequency changes; these changes limit fixed target cancellation. The klystron, on the other hand, is simply a power amplifier and has no influence on the frequency determining system. A reference oscillator (stable local oscillator) is maintained at an almost constant frequency and is effectively isolated from the output stage by intermediate amplifiers.

The reference oscillator output which is the input to the klystron amplifier is mixed with the coherent oscillator output to produce the desired output frequency.

Because the frequency determining oscillators operate at a low level, both can be made extremely stable and the pulse-to-pulse frequency variations are almost negligible. No AFC is needed in such a system and the pulse-to-pulse phase variations of the transmitter output stages are limited by providing a modulator pulse having a suitably flat-topped portion. The fixed target cancellation of an associated MTI system is considerably improved by the low pulse-to-pulse phase variations, and the radar transmitter is no longer considered the limiting factor in the MTI performance of a radar system.

The disadvantages of the klystron are chiefly the circuit complexities of the transmitter. A series of amplifiers are inserted between the frequency determining oscillators and the power amplifier klystron. These stages must be individually tuned when the operating frequency is changed. Most of these stages are pulsed, and synchronizing problems arise. Power supply requirements are also increased. The bulk of the equipment required in a high-power klystron radar system limits the mobility of such a system, and the radar installation employing klystrons is either a fixed station or one of limited mobility. Regardless of the disadvantages, the klystron system complexity and bulk are justified in that they insure consistent MTI performance which excels that possible with magnetrons.

### Klystron Mixer Tube

Another utilization of the multicavity klystron, other than a power amplifier, is that of a mixer. The klystron is used as a mixer tube primarily when the power amplifier klystron is used in the transmitting system. The mixer is used to mix the stable local oscillator (stalo) signal and the coherent oscillator (coho) signal to produce the transmitter frequency. A typical klystron mixer circuit is shown in Figure 12-69 and will be referred to as the synchrodyne tube.

The synchrodyne tube mixes the coho output with the stalo output to produce pulsed energy of the proper frequency for application to the

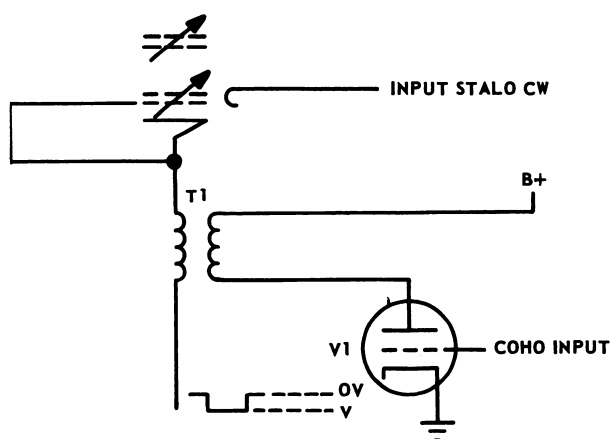


Figure 12-69. Simplified Klystron Mixer Circuit

first cavity of the power output klystron in the transmitter. The synchrodyne tube is a three-cavity, low-power klystron. The mixing procedure is effected in the flow of electrons in the drift tube from the cathode to the collector. Tuning of the cavities is accomplished by three tuning knobs (one for each cavity) which determine the resonant frequency of each cavity. These tuning knobs are accessible from the top of the tube.

The drive pulses applied to the cathode are negative; they are remodulated by the pulses of coho energy fed into modulation transformer T1 by driver tube V1. The drive pulse applied to the cathode causes beam current to flow for the entire period. But the r-f energy is applied for only a period during the drive pulse, and the mixing process is in operation during this time. The output of the stalo is applied to the first cavity of the synchrodyne tube continuously. The second and third (output) cavities of the synchrodyne tube klystron are so tuned that they are resonant only to a frequency that is the difference between the stalo and coho frequencies. Accordingly, the synchrodyne has an output only during the intervals when the coho pulses are applied to the synchrodyne tube's klystron cathode.

The mixing process is accomplished in the density modulated electron beam in the synchrodyne. The beam is modulated by the coho energy



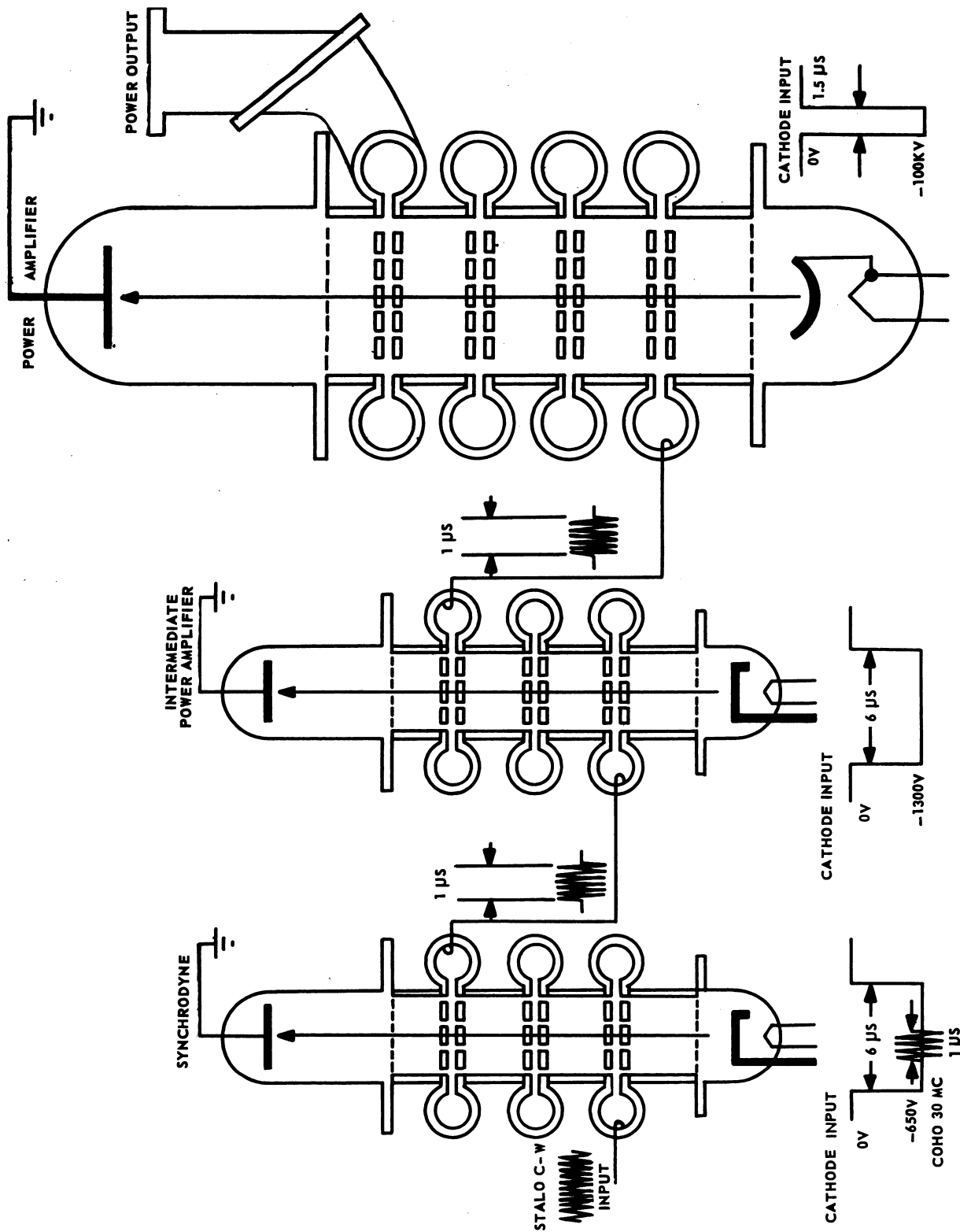


Figure 12-70. Multicavity Klystron Stages

and the first cavity is tuned to the frequency of the stalo. The second and third cavities are tuned to the difference between the coho and stalo frequencies. The collector of the synchrodyne tube is grounded since the beam voltage is applied as a negative pulse on the cathode. Figure 12-70 illustrates how the synchrodyne is used as a mixer tube in conjunction with an intermediate stage.

### TRAVELING-WAVE TUBE

The traveling-wave tube has, since the 1950's, become important to the circuit designer and the systems engineer in the field of microwave amplification. This does not mean that the traveling-wave tube is better than other microwave tubes in all respects. As yet it is somewhat inefficient when compared to magnetrons, although efficiencies of over 10% have been obtained. A related device, the "M" type *carcinotron* or magnetron amplifier, shows quite high efficiencies and essentially operates on the traveling-wave principle.

It is true that gains of 40 db have been achieved by the traveling-wave tube which can be rivaled only by the newer multiresonator klystrons. However, gain is not an all important consideration of the traveling-wave tube although gain is one of its major virtues. This factor, together with its low noise figure (some specially designed tubes have a noise figure of 6 db), make it an extremely useful device at microwave frequencies.

The simple structure of the traveling-wave tube has enabled manufacturers to build tubes that operate over a vast portion of the spectrum. By 1961 successful tubes had been built that operated as low as 300 mc and as high as 50,000 mc.

The major advantage of the traveling-wave tube is its tremendous bandwidth. It is this characteristic that makes it superior to other microwave devices. Bandwidths of one octave are not uncommon, and even greater bandwidths have been achieved. (A bandwidth of one octave is one in which the upper frequency is twice the lower frequency.) This feature allows the traveling-wave tube to be used in exploiting vast portions of the microwave spectrum.

### Characteristics

Most microwave amplifiers have a limited bandwidth. The bandwidth of a klystron, for example, is no greater than that of low-frequency pentodes. High-frequency triodes do exhibit a wider bandwidth than klystrons, but this is not significant when compared to the traveling-wave tube. Microwave triodes and klystrons also suffer the same limitation as do conventional tubes. As the bandwidth is widened, the gain is decreased to the point where no gain is available. Tube capacities and electron transit time also limit these tubes to relatively low microwave frequencies. (In the case of the triode, successful operation cannot be obtained much above 4000 mc.)

In the traveling-wave tube these limitations are overcome completely. There is no lumped input capacitance nor any resonant circuit. The tube is a transmission line with a negative attenuation in the forward direction and a positive attenuation in the backward direction. Negative attenuation is the exponential *increase* of voltage in the direction of propagation. Positive attenuation is the exponential *decrease* of reflected voltage. The bandwidth is limited to some extent by the input and output transducers connected to the tube, but these have been successfully made to cover a bandwidth greater than one octave. In some cases the bandwidth is so great that we have by no means fully exploited what is available.

In general, the traveling-wave tube compares very favorably with other microwave devices as to gain, noise figure, and simplicity of construction. Admittedly, it is not as efficient as the magnetron, although reasonable efficiencies have been obtained. Finally, it definitely provides amplification over a bandwidth commensurate with microwave frequency space available, and it is this feature that has caused it to be incorporated in radars, countermeasures equipment, repeaters and communications equipment, and radio astronomy devices.

### Construction

The traveling-wave tube was first announced in 1946. Since that time the large research organizations of the world have devoted much time

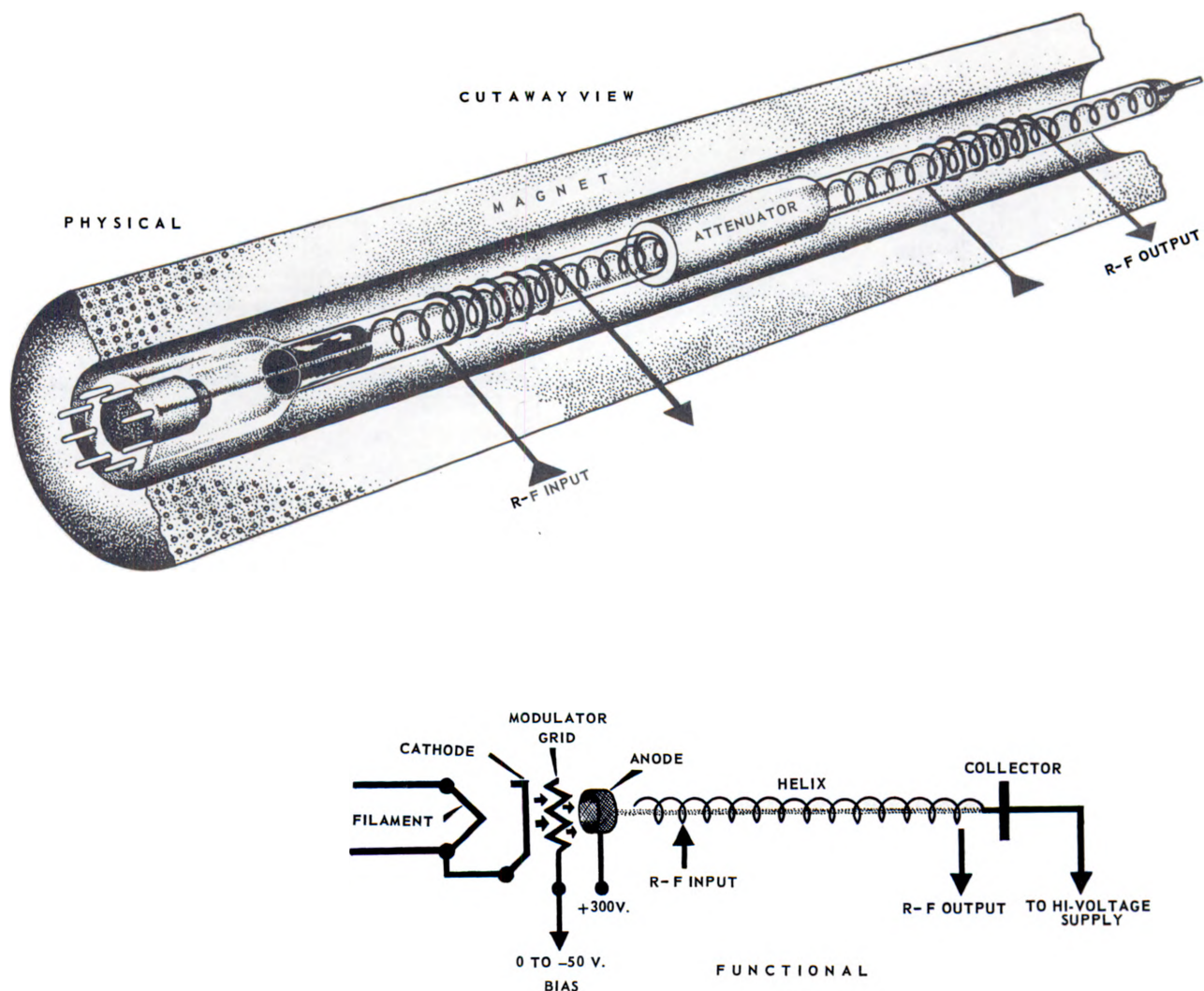


Figure 12-71. Traveling-Wave Tube

and money to perfect it. The traveling-wave tube consists of an electron gun that produces a stream of electrons which are kept in a tight beam with a magnetic focusing field supplied by a solenoid magnet. A cutaway view of the traveling-wave tube is shown in Figure 12-71. The electrons are accelerated by two different means. Initially they are accelerated by the specially designed electron gun that originally produced them. In the electron gun the heater boils off electrons from a parabolic-shaped cathode. This tends to focus the electron stream; small accelerator anodes increase the velocity. The result is a narrow beam that has the relative velocity of an electron accelerated by 1500 volts.

A positive potential on the helix and a collector

anode at the end of the tube continue to speed the electron flow. The electron beam is made to flow down the central axis of a long, loosely wound helix on its way to the collector. The beam is kept from spreading by the solenoid magnet surrounding the tube. The focusing system that is generally used is a long, wirewound, magnetic solenoid that completely surrounds the loosely wound helix portion of the tube. The axis of the solenoid magnetic field coincides with the central axis of the helix.

The ability of this magnetic lens to keep the electron beam focused into a tight area is a direct function of the current through the magnet. The higher the solenoid current, the tighter the beam. If the magnetic field of the solenoid were

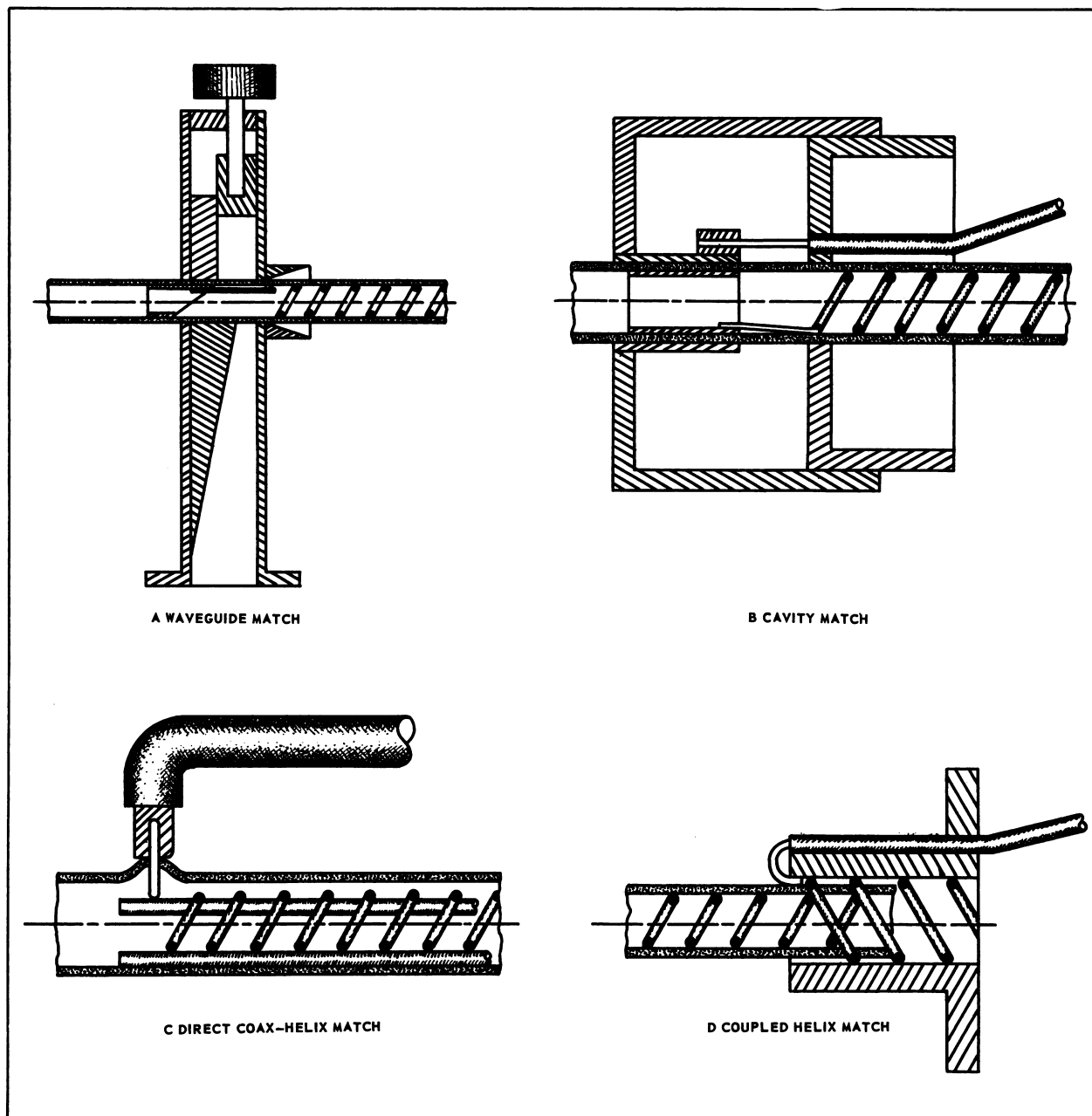


Figure 12-72. Traveling-Wave Tube Couplers

lost for an instant, the electron beam would spread, intersect the helix, ionize, and destroy the traveling-wave tube. The weight of the electromagnet and the associated power supply add to the bulk and complexity of traveling-wave tubes. In fact, this has prevented their use in some instances. An improvement has been sole-noids wound with aluminum foil instead of wire.

### Coupling

Some means must be provided to apply r-f energy to one end of the helix, and remove it from the other end. A transformer is one way of doing this. Another method is by waveguide matching (shown in Figure 12-72A). The waveguide is terminated in a nonreflecting impedance.

The helix is inserted into the waveguide. The coupling action is similar to that of a quarter-wave section. The efficiency of the system is rather good, but the waveguide has a far higher  $Q$  than the traveling-wave tube. This means that the broadband characteristics of the traveling-wave tube suffer in that the entire bandwidth is not available for amplification. The waveguide will not respond over such a wide spectrum.

The cavity match, illustrated in Figure 12-72B, is very similar to the waveguide match. Cavities can be made to resonate over wider ranges than waveguides, but they still have a high  $Q$  compared to the traveling-wave tube. The helix is placed at the mouth of the cavity ready to absorb any E-fields produced. The r-f is led to the cavity by coaxial cable and is introduced into the cavity by a quarter-wave section. A similar system removes the r-f from the other end of the helix.

The direct coax-helix match, shown in Figure 12-72C, is sometimes referred to as direct pin match. It is the simplest system of all. The center conductor of the input coax is connected directly to the helix of the traveling-wave tube. Although this is the method usually used in the high power traveling-wave tubes of today, it has some disadvantages. The high-voltage standing-wave ratio causes heating around the pin. Since this pin passes through the glass envelope on the way to the helix, the envelope is subject to heating and breaking at this point.

Figure 12-72D shows the coupled helix match, a widely used type of coupling. In this system the center of the coax input conductor is attached to a small helix. The major helix of the traveling-wave tube is inserted within this input helix where it acts as the secondary of a transformer. The input r-f energy is magnetically coupled to the main helix; the output r-f is removed by magnetic coupling also. Although the structure of a helix has good VSWR characteristics and is broader in bandwidth than cavities and waveguides, it is as yet mechanically unable to handle large amounts of power.

### Operation

To understand the operation of the traveling-wave tube, it will be helpful to compare it to that of the two-cavity klystron amplifier. In the

klystron an electron gun which is capable of great emission densities provides the electron beam that travels through input and output resonant cavities and is finally captured by a collector electrode.

The input cavity is tuned to the signal frequency to be amplified. Thus an r-f voltage is concentrated across the first gap. This alternating voltage has an E-field whose direction is parallel to the electron stream. Thus it will either add to or subtract from the velocity given to the electrons by the beam voltage. Electrons passing through the gap are either accelerated or decelerated. Therefore the electron beam, which has a uniform velocity given by the beam accelerating voltage, is also velocity-modulated at an r-f rate by the input r-f signal. Fast electrons catch up with slow electrons at specific distances beyond the input cavity to form bunches. The frequency at which these bunches of electrons arrive at points beyond the buncher cavity corresponds to the frequency of the input signal. One bunch will arrive during each cycle.

If the catcher cavity is tuned to the same frequency and placed at a point along the electron stream where bunches form, oscillations that follow the input signal will be induced in the cavity. There is high-power gain because the power required to bunch the beam is much less than the power that can be derived from it. The additional energy comes from the kinetic energy imparted to the beam by the DC accelerating voltage. The interaction between the electron stream and the r-f fields takes place only in the rather narrow gap. The  $Q$  of the cavities must be high to build up the r-f voltage across the gaps to high values for efficient interaction. This high  $Q$  cavity means that the klystron is a narrow band device.

The tremendous bandwidth of the traveling-wave tube compared to that of the klystron is the traveling-wave tube's major advantage. The traveling-wave tube differs from the klystron principally in that the electrons interact with a *traveling* wave, rather than with a *standing* wave. This interaction is distributed, rather than localized, as in a klystron. There are no high  $Q$  resonant circuits in a traveling-wave tube; therefore, amplification can occur over appreciably broad frequency ranges.

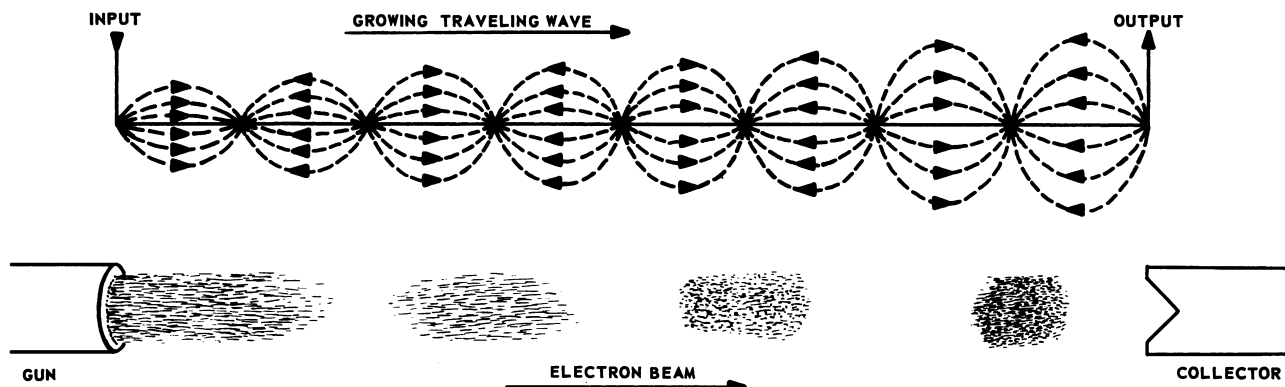


Figure 12-73. Simplified Traveling-Wave Concept

The simplest concept of a traveling wave is illustrated in Figure 12-73. An electron beam is passed along, or around, a transmission line consisting of a straight wire. This wire has an input and an output terminal. The electrons are gathered by a collector electrode, just as in the klystron. If the output is terminated in the characteristic impedance of the line, a wide range of radio frequencies will travel along the line; the line is nonresonant. Consider these frequencies. Part of their electric fields are parallel to the direction of travel of the wave and to the direction of the electron beam. This means that there will be interaction between the electrons in the beam and the electromagnetic wave on the transmission line wire. This interaction takes place the length of this hypothetical tube and for a wide range of frequencies.

If the electrons could be accelerated so that they traveled faster than the electromagnetic wave on the wire, they would become bunched due to the effect of the field parallel to the electron beam. Some of the electrons would be slowed down, while others would be accelerated. These bunches of electrons give part of their energy to the wave on the transmission line. As a result, the traveling wave increases in amplitude, as shown in Figure 12-73. Since the same action takes place for a wide range of frequencies, the traveling-wave tube has a very wide bandwidth.

However, this simple form of a traveling-wave tube will not work because waves travel along a straight wire at the speed of light. At present there is no known way of accelerating a beam of electrons to such speeds. For this reason travel-

ing-wave tubes are designed to use a transmission line that will delay the r-f wave so that the electron stream can be accelerated to the same or higher speeds.

A common type of delay line is a wire that is wound in the form of a long coil, or helix, as shown in Figure 12-74. The diameter and the number of turns per unit length are selected so that while the wave travels along the wire at the usual speed, its velocity along the axis of the helix is about one-tenth the velocity of light. Now it is possible to accelerate the electrons to the point where their speed matches that of the wave as it proceeds down the tube. The helical type of delay line works very well because there will be a large proportion of the electric field parallel to the electron beam. There will be longitudinal fields both inside and outside the transmission line (helix). Usually the beam of electrons is passed down the center of the coil because the field is more concentrated there.

If any of the signal reflected at the output end reaches the input, it will be greatly amplified. This would produce spurious oscillation. Figure 12-74 shows a lossy wire attenuator placed near the midpoint of the traveling-wave tube envelope. Another type of attenuator consists of a split cylinder of graphite which is placed around the coil near the center. These attenuators serve to isolate the input from the output, thus preventing oscillation. In the lower portion of Figure 12-74 is shown the traveling wave for the tube. Notice the loss in gain due to the attenuator. The bunches of electrons emerging from the attenuator build up a new c-w signal on the helix.



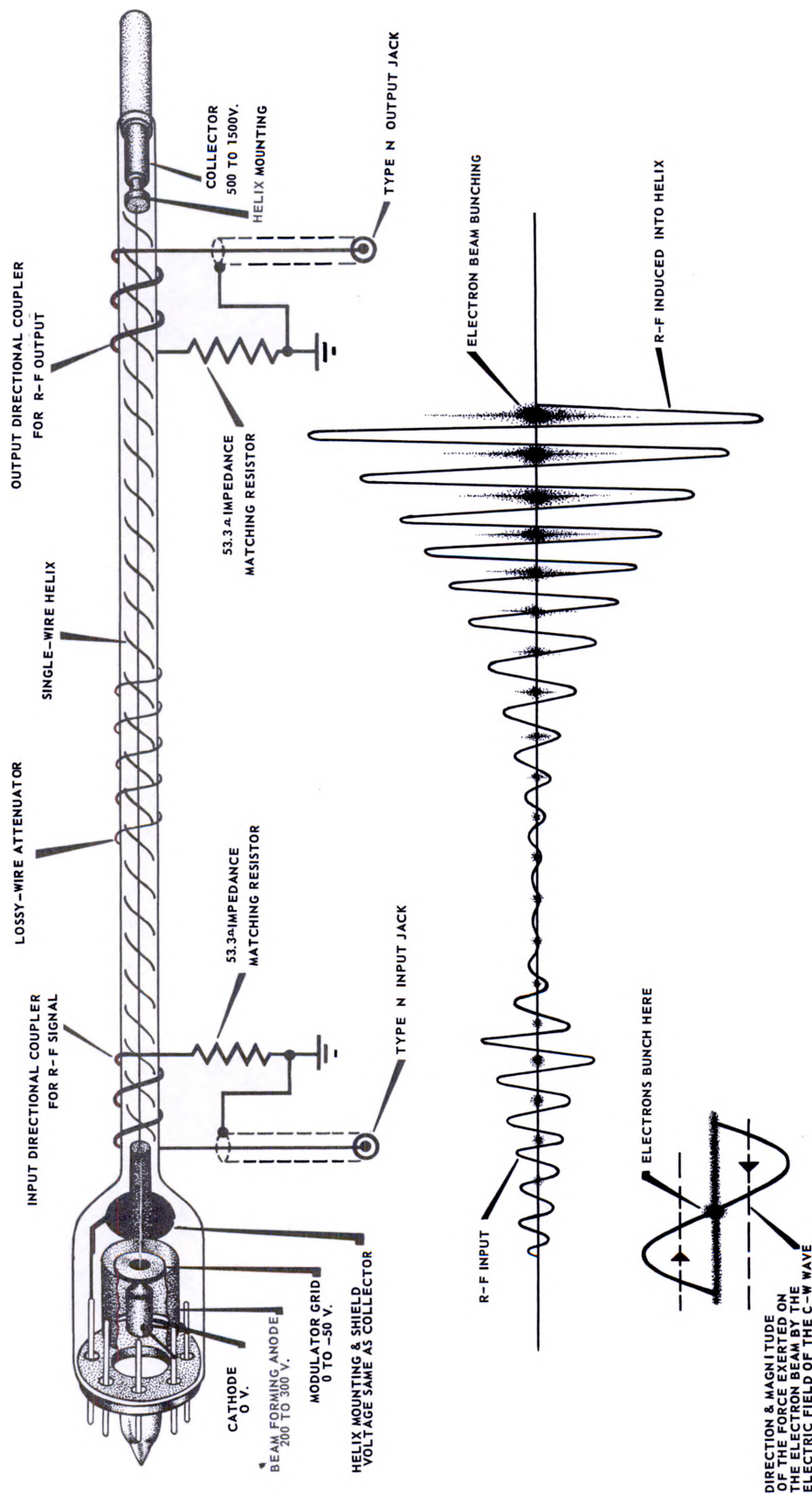


Figure 12-74. Functional Diagram of a Traveling-Wave Tube

This signal has the same frequency as the original input signal.

### Summary of Action in a Traveling-Wave Tube

In the typical low- and medium-power traveling-wave tubes the electron beam is directed through the center of a transmission line in the form of a coil, or helix. A continuous-wave signal of a predetermined frequency is coupled into the helix. Electrons are bunched due to the action of the fields. Amplification of the signal on the helix begins as the field formed by the bunches interacts with the electric field of the c-w signal. Each newly formed electron bunch adds a small amount of energy to the c-w signal on the helix. The slightly amplified c-w signal then causes a denser electron bunch which, in turn, adds a still greater voltage to the c-w signal. This process is continuous as the wave and the bunches progress down the helix.

Amplification increases as the greater velocity of the electron beam and the retarding effect of the negative points in the wave pull the electron bunches more nearly in phase with the electric field of the c-w signal. The additive effect of the two fields, exactly in phase, produces the greatest resultant of amplification. Attenuators placed near the center of the helix reduce all the waves traveling along the helix to approximately zero. This prevents reflected waves from returning to the tube input and causing spurious oscillations. The electron bunches pass through the attenuator unaffected. The electron bunches, after emerging from the attenuator, induce a new c-w signal on the helix. This new c-w signal has the same frequency as the original input signal. The electric field of the newly induced c-w signal interacts with bunched electrons to begin afresh the amplification process.

For a short distance, the velocity of the electron bunches is reduced appreciably due to the large amount of energy absorbed by the formation of the new c-w signal. Amplification increases as the greater velocity of the electron beam and the E-fields of the wave pull the electron bunches more nearly in phase with the electric field of the newly induced c-w signal. At a point of predetermined power, the amplified c-w signal is coupled out of the helix. Note that the amplified c-w signal is a "new" signal whose energy is completely supplied by the bunched electron beam.

### Types of Traveling-Wave Tubes

There are three basic types of traveling-wave tubes: the low level, the driver, and the power amplifier. The helix tied to the collector and separated from the accelerating anode is a distinguishing characteristic of the low level type. The driver traveling-wave tube has the helix separated from the collector and connected to the accelerating anode. The power amplifier traveling-wave tube has the collector, helix, and accelerating anode all separate and at the same potential.

### Advantages and Disadvantages

The traveling-wave tube is used because of its very wide bandwidth characteristics. This unique advantage is almost offset by several disadvantages. The traveling-wave tubes in use today must rely on external magnetic focusing of the electron beam. For the S-band this amounts to a solenoid current of 11 amperes with a voltage drop across the solenoid of approximately 15 to 85 volts DC. None of this power expenditure can be recovered in the output.

The traveling-wave tube is also a low efficiency device. The lost power is in the form of heat, which calls for a high volume of air conditioning or liquid cooling. Should a power amplifier traveling-wave tube be turned on without forced cooling, the tube would fail almost immediately. Even slight heating can deform the helix, and a nonuniform helix causes "holes," or areas of no transmission, in the output. Holes that exist in normal traveling-wave tubes are due to errors in the manufacturing process.

Because these holes are known to exist, their position (frequency) must be accurately determined. This leads to another disadvantage in that a traveling-wave tube must be thoroughly checked out before it can be placed in its intended equipment. Perhaps one of the most elaborate of tube testers has been constructed for traveling-wave tube checkout. Such a piece of equipment must always be available when equipment using these tubes is to be serviced. Each traveling-wave tube must be tested and evaluated independently. At this stage of development, it is impossible to substitute one traveling-wave tube for another. So much difference exists among them that they must be

calibrated separately. Then, only after a comparison has been made, can one be substituted for another of the same type.

## BACKWARD-WAVE OSCILLATOR

The backward-wave oscillator provides a flexible source of microwave energy that can be voltage-tuned to produce a frequency range as large as 2:1. The output frequency of the backward-wave oscillator is determined by a frequency-selective feedback and amplification process rather than by resonant circuits as used in conventional microwave oscillators.

### Structure of Helix Type

The backward-wave oscillator tube consists of an electron gun, a helix structure, and a collector at the far end of the helix. Physically, the backward-wave oscillator resembles the traveling-wave amplifier tube, although for comparable frequencies it is larger in diameter and somewhat shorter in length. Another difference, not apparent from a visual inspection of the tube, is that the helical backward-wave oscillator uses a hollow electron beam with a strong concentration of the electrons near the helix. This hollow electron beam is focused along the length of the helix by a strong magnetic field supplied by an axial solenoid surrounding the tube.

The r-f output of the backward-wave oscillator is a result of the interaction between the electron beam and the electric fields accompanying a microwave signal present on the helix. The term backward-wave oscillator is quite appropriate for this tube since the r-f energy moves and builds up in a direction opposite to that of the electron beam and is coupled out at the gun end of the tube via the helix terminal.

### Theory of Operation

The operation of the backward-wave oscillator tube may be explained in terms of a series of feedback loops similar to those common to low-frequency electronic circuits. Each of these regenerative loops can function as an amplifier or an oscillator and is designed so that the phase shift around the loop is one cycle. One of these feedback loops is shown in Figure 12-75 where, using conventional circuit terminology, the for-

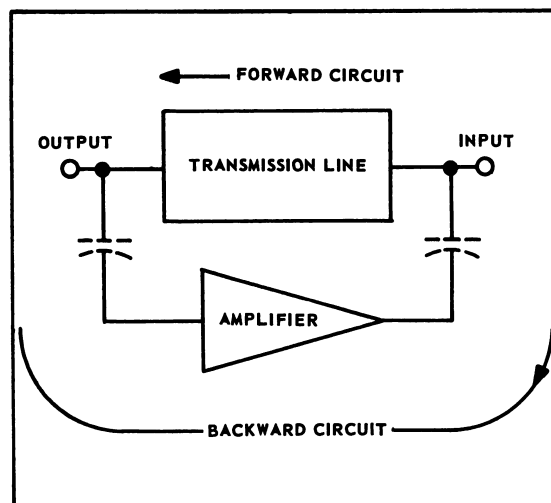


Figure 12-75. Typical Feedback Loop

ward circuit consists of a section of transmission line and the backward circuit is a unilateral amplifier connecting the output of the transmission line to the input.

In this circuit positive feedback will occur when the amplifier gain becomes sufficiently high to overcome the loss in the transmission line. The loop will oscillate at a frequency for which the total phase delay is one or more cycles. If the amplifier is designed for limited high-frequency response, oscillations will occur only when the phase delay is one cycle. The frequency of oscillation can be shifted by changing the phase delay in the amplifier. The essential feature of the voltage-tuned backward-wave oscillator is that the frequency of oscillation can be changed electrically by changing the phase delay in the amplifier.

**FEEDBACK LOOPS.** Figure 12-76 shows a chain of identical regenerative feedback loops. Along the top of the chain is a series of transmission line sections that will support a wave moving either to the right or to the left. Along the bottom of the chain is a series of unilateral amplifiers in which signals can pass only in the left-to-right direction. Each loop consists of a transmission line, coupling capacitance and an amplifier transmitting from left to right. In operation, positive feedback, which leads to regenerative amplification and oscillation, makes use of a

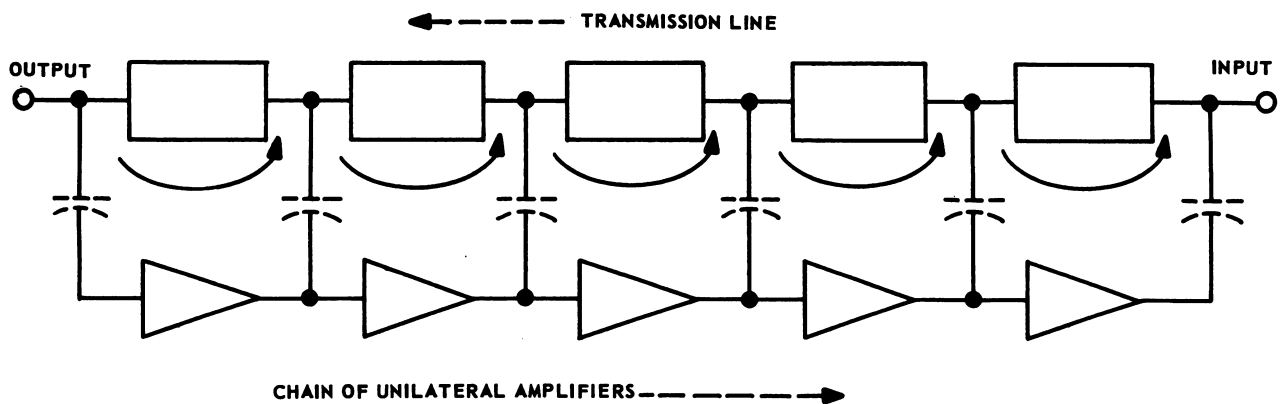


Figure 12-76. Typical Chain of Feedback Loops

wave going from right to left on the transmission line when the phase delay in a single loop is just one cycle. The total phase delay around a group of  $n$  loops then will be  $n$  cycles.

For low values of amplification, the chain of loops will act as a regenerative amplifier operating at the frequency which provides positive feedback. But if the transmission line is terminated in its characteristic impedance at the input and the amount of amplification is increased, oscillations will start. The frequency of oscillation is controlled by the phase delay in the amplifier chain.

With this background, we can now examine the actual functioning of the backward-wave oscillator tube. Figure 12-77 shows a cross section of the helix, and a portion of the electron beam. The helix structure consists of a cylindrically-wound flat-wire tape; the electron beam is hollow

and passes very close to the helix turns. The strong axial magnetic field focuses the electrons in the beam and allows movement only in the direction of the axis of the tube. The lines of force of the electric fields associated with an r-f wave traveling along the helix are also shown in Figure 12-77. Although these fields rotate around the helix at the velocity of light, the axial velocity is much slower, depending on the ratio of the turn-to-turn spacing of the helix to its circumference. The axial velocity of the r-f wave is approximately equal to the velocity of the beam wave.

The lengthwise electric fields are strong between helix turns and very weak directly under the wire itself (see Figure 12-77), since the electric fields cannot exist parallel to a conductor. The field between the helix turns has a great effect on the velocity of the electrons in

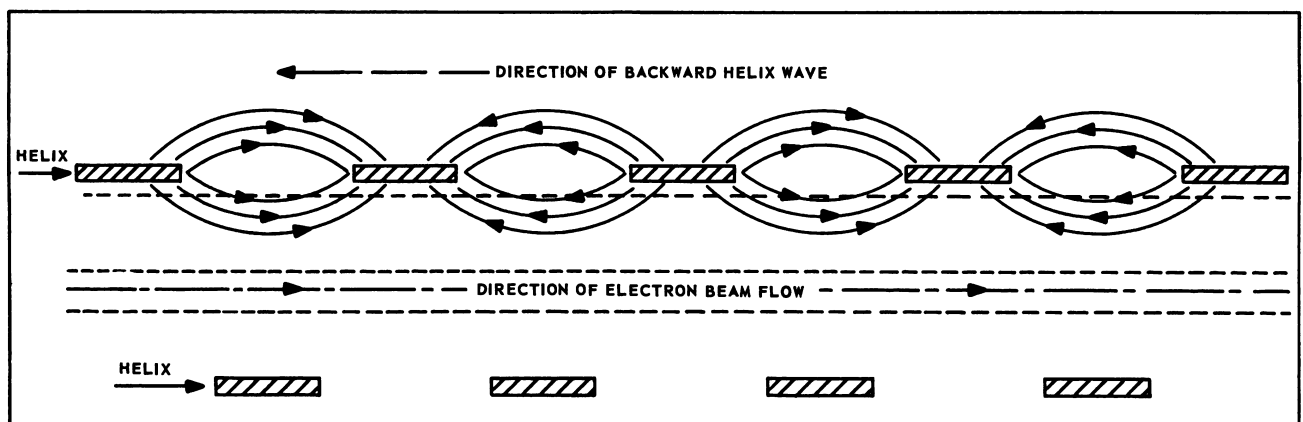


Figure 12-77. Cross Section of Helix, Electron Beam and Helix Wave

the beam, producing an interaction process. This process is represented by the capacitive coupling between the transmission line and the amplifier chain shown in Figure 12-76. In this way feedback loops are formed between the midpoints of adjacent helix gaps.

Although the concept of individual feedback loops is a useful device for explanation, the backward wave interaction is actually a continuous process. The maximum coupling between the helix wave and the beam occurs midway between gaps and gradually tapers off to a minimum directly under the helix turns. One of these regenerative loop chains exists at each angular position around the helix. Each of these regenerative loop chains is independently coupled to the helix transmission line so that the net effect is a continuous amplification and feedback process occurring down the entire length of the tube.

**AMPLIFICATION.** The basic mechanism of amplification is a velocity modulation process that causes the electrons to bunch in the beam. Figure 12-78 shows the sinusoidal variations in amplitude of the electric field at the midpoint between helix turns. The phase relationship between the backward wave on the helix and the velocity of the electron beam is such that each specific portion of the electron beam will be affected by an electric field of the same phase as it passes successive gaps down the helix.

An electron at point A in Figure 12-78 experiences a continuous decelerating effect caused by

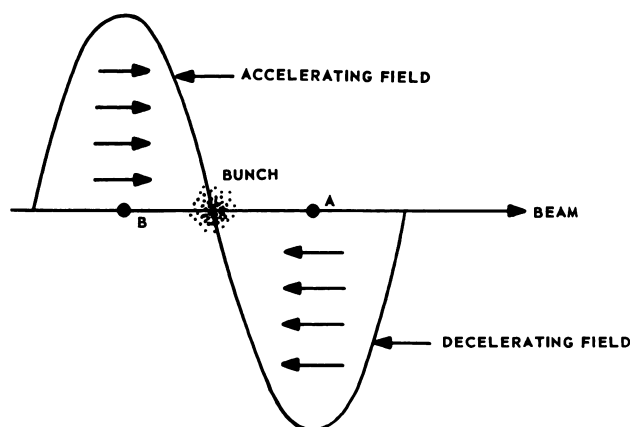


**Figure 12-78. Axial Electric Field of Helix Wave Provides Velocity Modulation and Bunching of Electron Beam**

fields of the same phase and direction of force as it proceeds down the tube. In a like manner an electron at point B will be continuously accelerated in its journey down the tube. In this way some parts of the electron beam are slowed down while others are advanced. The net effect is a bunch formed at the midpoint between the accelerating and the decelerating fields (Figure 12-78). Since a bunch will be formed in the beam for each wave on the helix, 20 to 40 bunches will be traveling along the beam (each surrounded by a wave on the helix) at any given instant.

At this point it should be mentioned that the average electron velocity of the beam is slightly faster than the effective phase velocity of the amplifier chain. This means that the electron bunches will advance a quarter of a cycle as they approach the collector end of the tube. Thus they encounter the full decelerating effects of the electric field and give up a maximum amount of kinetic energy to the wave on the helix.

Figure 12-79A shows that the density of the electron bunches increases according to an exponential relationship. Figure 12-79A shows the envelope of the bunching, rather than instantaneous amplitudes, since many r-f cycles exist

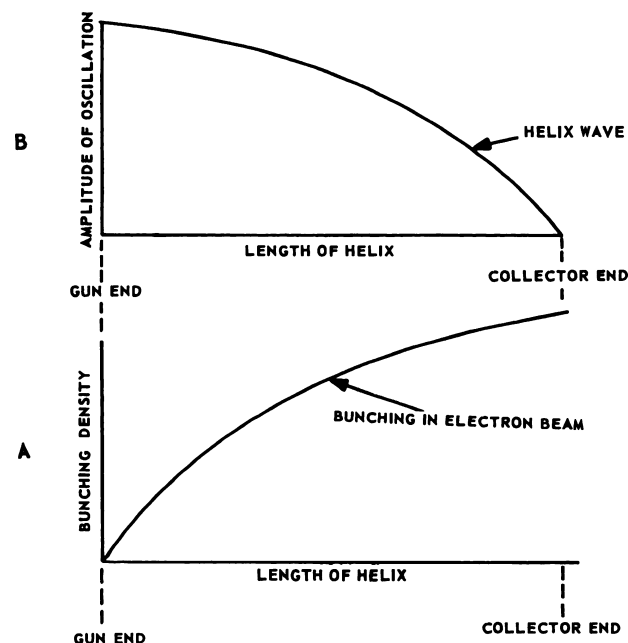


**Figure 12-79. Relative Amplitude of the Helix Wave and Electron Bunching Depicted in Envelope Form**

along the length of the backward-wave oscillator tube. The wave on the helix moves from right to left toward the gun end of the tube and gains amplitude according to the degree of electron bunching in the beam. In this way the envelope of the wave on the helix shown in Figure 12-79B is the result of the bunching envelope so the maximum energy transfer from the beam to the wave on the helix occurs at the collector end of the tube.

**BEAM VELOCITY CONTROLS FREQUENCY.** Now that a correspondence between the chain of lumped regenerative loops and the helix backward-wave oscillator tube has been established, it can be seen that if the velocity of the electron beam is varied, the phase delay around each of the regenerative loops will be changed. If the electron beam is at the correct velocity, the chain of regenerative loops will oscillate at a frequency where the phase delay of each loop is equal to one cycle.

Oscillations begin in the backward-wave oscillator in much the same way as they begin in other oscillators. Noise waves are established on the helix from the shot noise coupled from the electron beam and from the thermal energy developed in the termination at the input end of the tube. The waves traveling backward on the

tube (to the left) velocity modulate the beam. Velocity modulation causes the electrons to bunch and, in turn, reinforce the wave that exists on the helix at the frequency where the single-loop delay is equal to one cycle. In this way oscillations are built up at a single frequency determined solely by the electron beam velocity, which is a function of the cathode-to-helix voltage of the tube.

### **Crossed-Field Oscillator**

There has been extensive research and development in a newer oscillator called the crossed-field oscillator. This type of oscillator provides many advantages over other types of existing backward- and forward-wave oscillators. In comparison with the helix backward-wave oscillator, the crossed-field oscillator offers higher average c-w power, higher efficiency, broader bandwidth, and smaller size and weight per unit of power output. The crossed-field oscillator uses a permanent magnet with the advantage of elimination of solenoid power requirements and heat generation.

The principle of operation of the crossed-field oscillator is the same as that of the traveling-wave tube with the exception that the basic electron interaction is of magnetron interaction rather than that of klystron interaction.





# Antennas

This chapter deals with the basic principles of antennas. It discusses antennas in general, the principles of electromagnetic radiation and its application to radar antennas, various antenna arrays, and typical airborne radar antenna systems.

## FUNCTION OF THE ANTENNA

An antenna can be defined as a conductor or system of conductors used either for radiating electromagnetic energy into space or for collecting electromagnetic energy from space. To have adequate signal strength at the receiver, either the power transmitted must be extremely high, or the efficiency of the transmitting and receiving antennas must be high because of the high losses in wave travel between the transmitter and the receiver, and the lower sensitivity of the ultra-high frequency range receivers. Since it is difficult to generate large amounts of power efficiently in the higher frequency ranges, it is especially important that as much of the available signal at the transmitter be converted into radiated energy as possible, and that as much of the radiated energy as possible be picked up at the receiver.

## ANTENNA RECIPROCITY

Fortunately, separate antennas seldom are required for transmitting and receiving radio energy. Any antenna transfers energy from space to its input terminals with the same efficiency

with which it transfers energy from the output terminals into space, assuming, of course, that the frequency is the same. This property of interchangeability of the same antenna for transmitting and receiving operations is known as *antenna reciprocity*. Antenna reciprocity is possible chiefly because antenna characteristics are essentially the same regardless of whether an antenna is sending or receiving electromagnetic energy. Because of antenna reciprocity, most radar sets use the same antenna both for receiving and for transmitting.

An automatic switch in the radio frequency line first connects the antenna to the transmitter, then to the receiver, depending upon the sequence of operation. Because of reciprocity of radar antennas, this chapter treats antennas from the viewpoint of the transmitting antenna with the understanding that the same principles apply equally well when the antennas are used for receiving electromagnetic energy.

## DIRECTIONAL PROPERTIES

Usually, the most important characteristic of a radar antenna is its directional property, or its directivity. Directivity means that an antenna radiates more energy in one direction than in another. In radar operations some antennas are required to send all energy in one direction in order that as much as possible of the electromagnetic energy generated by the transmitter strike an object in a given direction. In other systems it is desirable for the energy to be radiated equally well in all directions from the

source. An example of an antenna system that radiates energy in a given direction is that of an airborne navigation and bombing set. In this set there is a limited amount of power available at the transmitter. To achieve maximum benefit from this minimum power, all of it is sent in the same direction.

Since the antenna in this set is also used for reception, it likewise receives electromagnetic energy from only one direction. Because of design features, it is possible to tell the direction of an object at which this directional antenna is sending energy or the direction of the object from which the antenna is receiving energy. Furthermore, the physical position of the antenna is indicative of the direction of the object. An example of a nondirectional radar antenna is the antenna installed in a radar beacon. This antenna must receive energy equally well from all directions in order that radar-equipped aircraft can ascertain position regardless of direction from the beacon antenna.

### ANTENNA EFFICIENCY AND GAIN

An efficient antenna is one which wastes very little energy. The higher the antenna efficiency, the less the energy loss. The greatest cause of decreased antenna efficiency during the conversion of guided waves to free-space waves, is the loss of energy in the form of heat. Heat losses will be discussed later. Actually, antenna efficiency may be high; in a single antenna element the efficiency may be as great as 90%.

An entire antenna system, properly designed, can concentrate all its radiated energy in a given direction with virtually none being lost in other directions. The gain of the antenna is the ratio of the maximum radiation intensity of the antenna in question to the maximum radiation intensity from a reference antenna with the same power input. The gain of a directional antenna compared to a nondirectional, or isotropic, antenna can be measured by measuring the intensity of the radiated field on the principal axis at a known distance from the antenna.

The field intensity that would exist with the isotropic radiator with that same power input is calculated. The ratio of the intensity of the directional antenna to the intensity of the nondirectional antenna is the gain. The effectiveness of a directional antenna in a particular di-

rection compared to a standard dipole is also known as antenna gain. It is usually expressed as the ratio of the standard antenna power to the directional antenna power that will produce the same field strength in the desired direction.

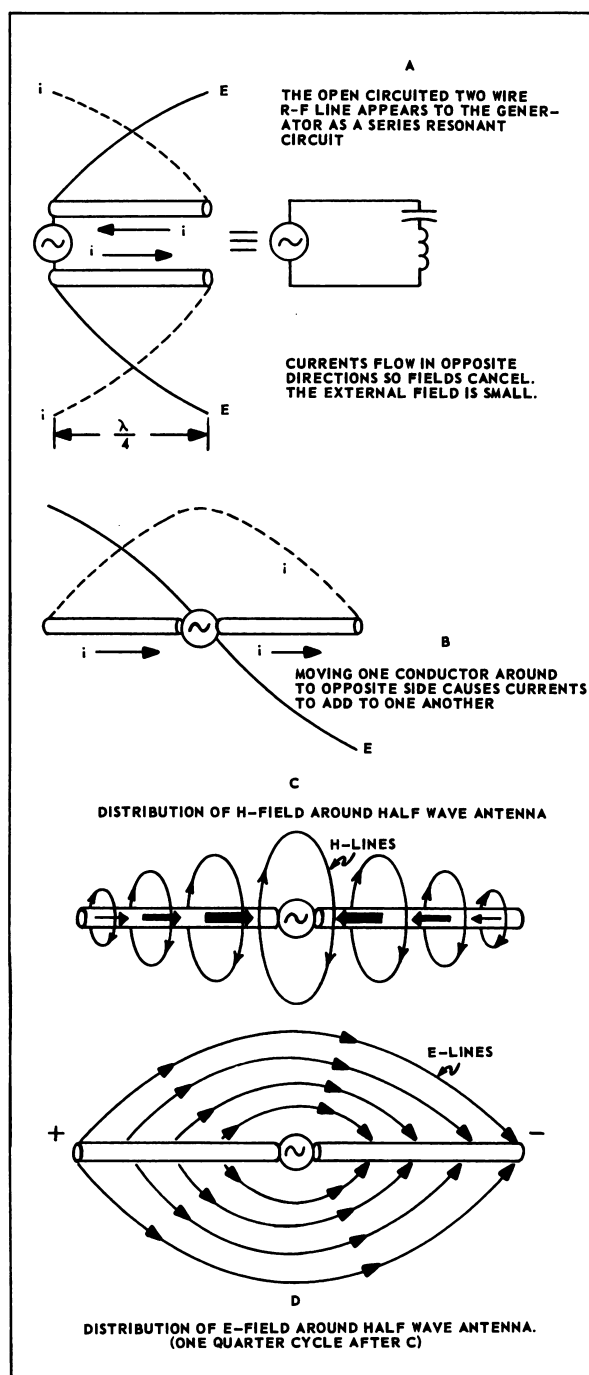


Figure 13-1. Development of Half-Wave Antenna from Open-End Quarter-Wave R-F Line

## ELECTROMAGNETIC ENERGY IN SPACE

Because the half-wave dipole antenna is the fundamental element in an antenna system, it will be used as a starting point for discussing the radiation of electromagnetic energy into space. A dipole antenna can be developed from an open-end transmission line one-quarter wavelength long. Figure 13-1A shows this quarter-wave section and the standing waves of current and voltage developed on the line. The current flow, as indicated, is flowing in opposite directions in the two conductors. The magnetic fields due to these currents cancel. The electrostatic fields, too, are in opposite directions.

Efficient radiation requires a very heavy field density; therefore, the fields must reinforce, or aid, each other. Suppose the ends of the two conductors are spread apart. The direction of current flow doesn't change, but the magnetic field is being "opened" into space, and less cancellation occurs. The electrostatic field is opened out instead of being concentrated in the small space between the two conductors. If the two conductors are moved until they are in line with each other as shown in Figure 13-1B, the current flow in the two conductors is now in

the same direction. The standing waves are exactly as they were before.

The magnetic (H) field about the wire is shown in Figure 13-1C. The direction of current flow is indicated by the arrows, from left to right. The magnetic field is in the direction indicated. Since maximum change of current takes place at the center of the dipole, the magnetic field is greatest at the center. Figure 13-1D shows the electrostatic (E) field about the conductor when the difference in potential between the ends of the conductor is the greatest. This does not occur at the same time as the maximum magnetic field. Since the voltage and current on the dipole are  $90^\circ$  out of phase, the electrostatic and magnetic fields surrounding the antenna are also  $90^\circ$  out of phase.

Figure 13-2 shows the relationship between the variations in field intensities. When maximum current is flowing, there is no difference in potential between the ends of the conductors. The magnetic field about the conductor is maximum (Figure 13-2A). As the current continues to flow, a charge is built up on the conductor and the current decreases. The charge built up increases and the electrostatic field which accompanies the charge also builds up. The current flow

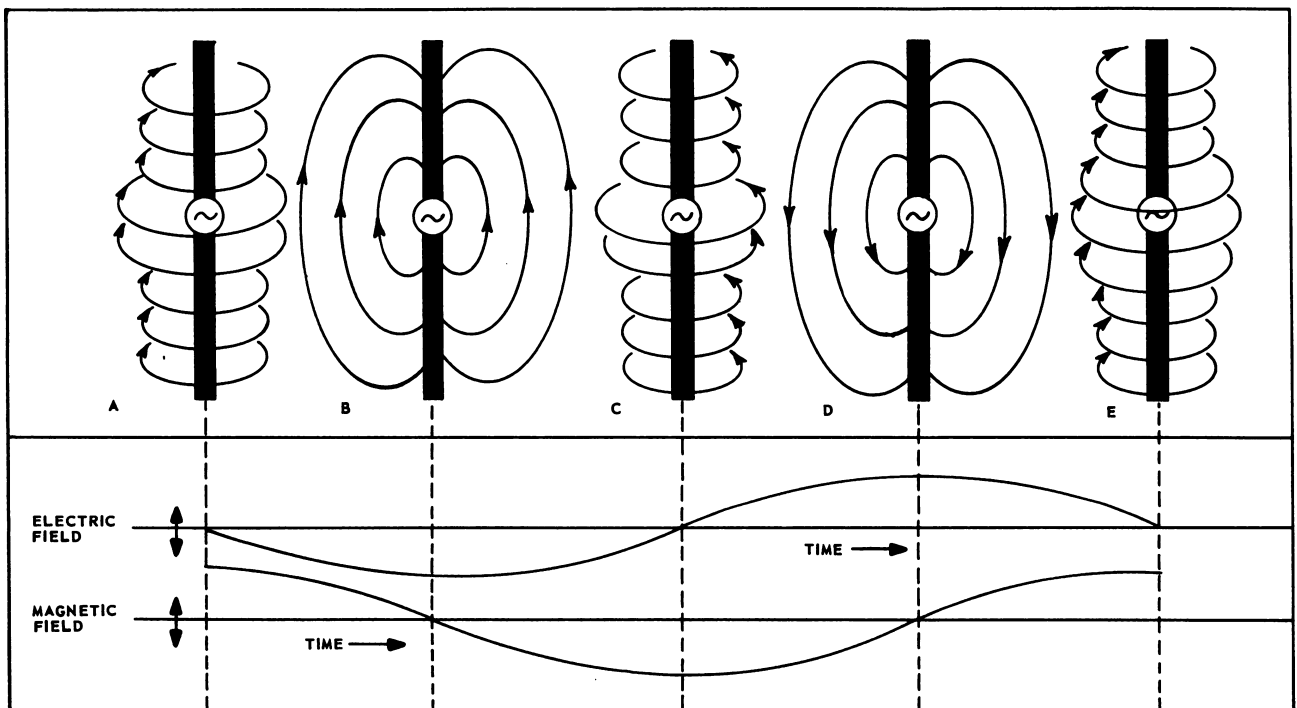


Figure 13-2. Variations in Field Intensities

and its accompanying magnetic field decrease. When the charge has built up to maximum, the current flow has decreased to zero. The electrostatic (or electric) field is at its maximum intensity (Figure 13-2B). When the current starts to flow in the opposite direction through the conductor, the magnetic field builds up as the electric field decreases. Figure 13-2C shows the fields when current is maximum in the direction opposite to A.

Figure 13-2D shows the electric field at a maximum when the preceding current flow has stopped momentarily and the magnetic field is zero. In E there is no charge and the flow of current is again maximum. Notice that the magnetic and electrostatic lines of force are at right angles to each other. In addition, these fields reach their maximum intensity a quarter of a cycle, or 90 electrical degrees, apart. It is said that these two fields are in *space and time quadrature*. Quadrature is the term used to express the phase relationship between two periodic quantities of the same period when the phase difference between them is one-fourth of a period. The E and H fields are 90° out of phase (time) and the lines of force are at right angles to each other (space).

The fields associated with the energy stored in the resonant circuit (antenna) are called the *induction* fields. These fields decrease with the square of the distance from the antenna. They are only local in effect and play no part in the transmission of electromagnetic energy. They represent only the stored energy in the antenna and are responsible only for the resonant effects that the antenna reflects to the generator. The

fields set up in the transfer of energy through space are known as the *radiation* fields. Although these fields decrease with distance from the antenna, this decrease is much less rapid than the decrease of the induction fields. This is because the decrease is linear. Therefore, the radiation fields reach great distances from the antenna. It is the radiation fields that are responsible for electromagnetic radiation.

### Mechanics of Radiation

There are many theories as to the why and how of radiation. While these theories are responsible for the advance in antenna design, they do not provide a simple and clear physical picture of the mechanics of radiation. The simplified physical picture does not tell the entire story. But the important fact to remember is that radiation does take place.

Figure 13-3 gives a simple picture of an E-field detaching itself from an antenna. The H-field will not be considered, although it is present. In A the voltage is maximum and the electric field has maximum intensity. The lines of force begin at the end of the antenna that is positively charged and extend to the end of the antenna that is negatively charged. Note that the outer E-lines are stretched away from the inner ones. This is due to the repelling force between lines of force in the same direction. As the voltage drops, the separated charges come together, and the ends of the lines move toward the center of the dipole (Figure 13-3B). But, since lines of force in the same direction repel each other, the centers of the lines are still being held out.

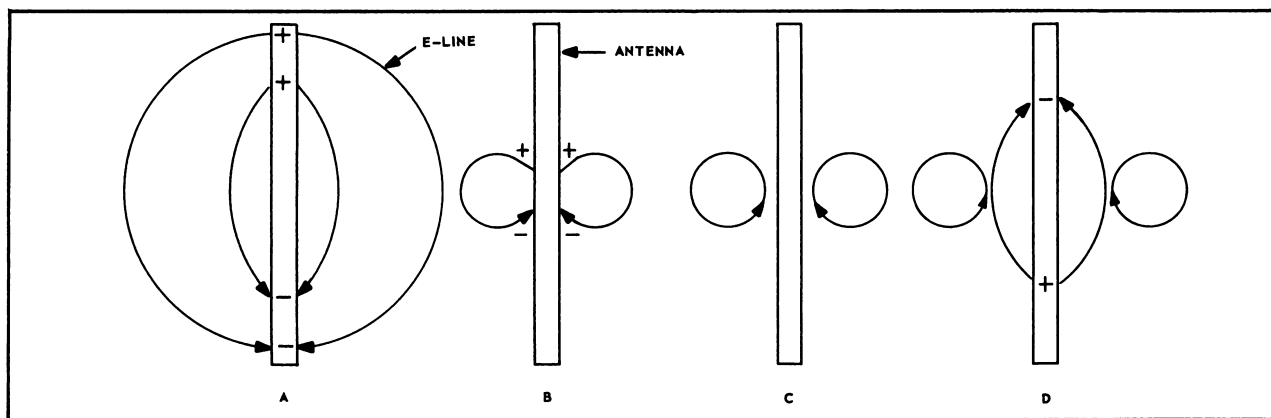


Figure 13-3. E-Field Detaching Itself from an Antenna

As the voltage approaches zero, some of the lines collapse back into the dipole, but in others the ends of the line come together and form a complete loop. Notice the direction of these lines of force next to the dipole in Figure 13-3C. At this point the voltage on the antenna is zero. As the charge starts to build up in the opposite direction, electric lines of force again begin at the positive end of the antenna and stretch to the negative end of the antenna. These lines of force, being in the same direction as the sides of the closed loops next to the antenna, repel the closed loops and force them out into space at the speed of light. As they travel, they generate a magnetic field in phase with them.

Another factor to remember about radiation is that the ease with which it may occur varies with frequency. At lower frequencies, such as 60-cycle power frequency for example, voltage on the antenna changes so slowly that the component of energy radiated is extremely small and is of no

practical value. At higher frequencies, such as 50,000 cps and up, the radiated energy is great enough to meet communications requirements.

### Fields in Space

Figure 13-4 shows the manner in which radiation fields are propagated from the antenna. The E-field and the H-field are shown as separate sets of flux lines about the antenna. The electric flux lines are the closed loops on each side of the antenna. The magnetic flux lines are closed circular loops which have their axis around the antenna, or in other words, the antenna is the center of each loop. They are represented as dots and crosses. The sine waves that are labeled the curves of radial variation of flux density indicate the relative field strength at various distances and angles from the antenna. Since the field configuration is not a standing wave, but a traveling wave phenomenon, the magnetic and electric

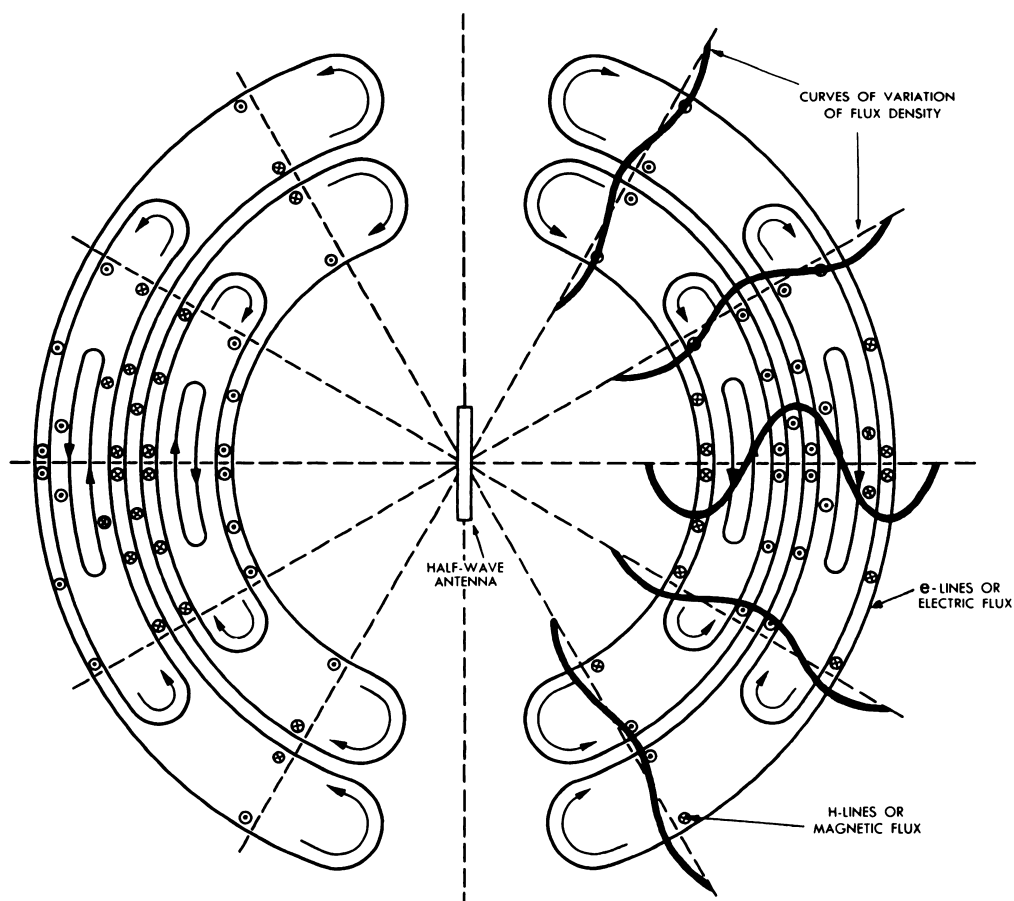


Figure 13-4. Fields in Space Around a Half-Wave Antenna

fields are in phase and thus the sine waves apply to the flux density of either field.

In the direction perpendicular to the antenna, both the H-field and the E-field are strong, for this is the direction where both fields originally formed. In the direction parallel to the antenna, or off the ends, no H-field forms at all and only a very small E-field. The flux density, therefore, is small in these two directions. Due to the directional properties of the half-wave antenna, most of the radiated energy travels in the direction perpendicular to the antenna, while very little energy travels in the direction along its axis.

As previously stated in the discussion of waveguides, no moving E-field can exist without an H-field, and no H-field can be propagated without an associated E-field. Similarly, with the propagation of electromagnetic energy into space, no moving electrostatic forces can exist without magnetic stresses existing in space, and no magnetic force in motion can exist without an electrostatic stress. Each creates the other, and one cannot exist without the other.

The direction of motion is the travel of the fields away from the antenna. The current associated with the magnetic field does not flow because in space there is no conductor to carry current, but the field exists, nevertheless. To visualize a field existing without current flow, think of a moving magnetic field cutting a glass rod. A voltage (electrostatic stress) is induced in

the rod, but there is little actual electron movement because the rod is a good insulator. Magnetic lines move in space and set up electric stresses in space in a similar manner.

### Propagation into Space

Any part of the electromagnetic field set up by an antenna travels away from the antenna in a straight line. There are many parts to this field and many directions in which energy travels. Shown in Figure 13-5 are four of the large number of paths that the energy can take. These and other paths that energy takes affect reception of radiated energy. If there is nothing between the emitting antenna and the receiving station ( $R_1$ ), some energy will travel directly to  $R_1$  via path 3. Receiving station  $R_2$  cannot receive energy because the earth is between the two points and because the energy cannot go through the earth for any distance. Some of the energy will follow path 2 out into space.

At some height above the earth there is a heavily-ionized layer of atmosphere, called the *ionosphere* (refer to Chapter 15). This constitutes a change of media through which the energy must travel, and is sufficient to refract a wave. In the case of path 2 the refraction is sufficient to bend the energy wave out of the ionosphere back to the earth. Therefore, receiving station  $R_3$ , which is beyond the horizon and not located for

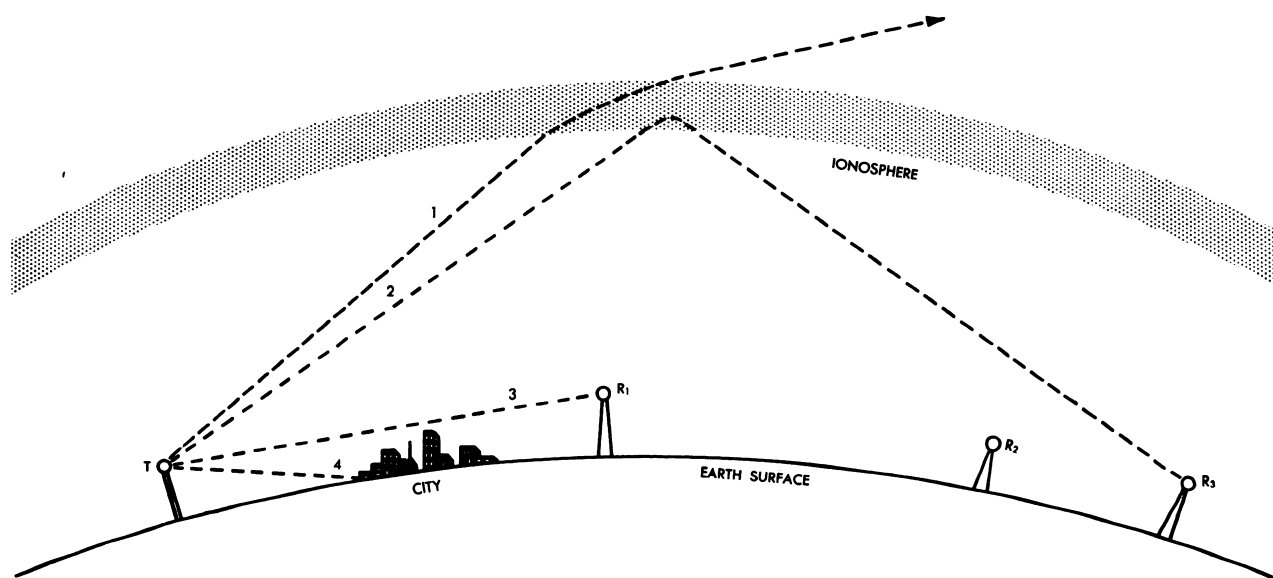


Figure 13-5. Effectiveness of Energy Traveling in Various Directions from an Antenna



receiving a direct ray, receives the reflected ray from the ionosphere.

It is the ionosphere that makes possible round-the-world communications. If the angle of incidence of an energy wave with the ionosphere is too great, the angle of refraction may not be sufficient to bend the energy back toward the earth and it may go on through the ionosphere and be lost, as shown by path 1. Paths 3 and 4 follow along the surface of the earth and are called *ground waves*. The waves bent back to the earth by the ionosphere are called *sky waves*.

The angle of refraction depends greatly upon the frequency of the wave. The higher the frequency, the less bending of the wave.

### THE HALF-WAVE ANTENNA

As stated previously in the discussion of electromagnetic radiation, a half-wavelength conductor is the simplest of the radiating elements. Considerable radiation occurs in this element because of its resonant characteristics and its ability to store large amounts of energy in induction fields. Resonance causes high voltages and high circulating currents and they, in turn, produce strong fields around the antenna.

In Figure 13-6 you can see that the voltage standing wave is high at the ends of the antenna and low at the center. As previously explained, this is also the case with the quarter-wave, open-circuit, two-wire line, from which the half-wave antenna is developed.

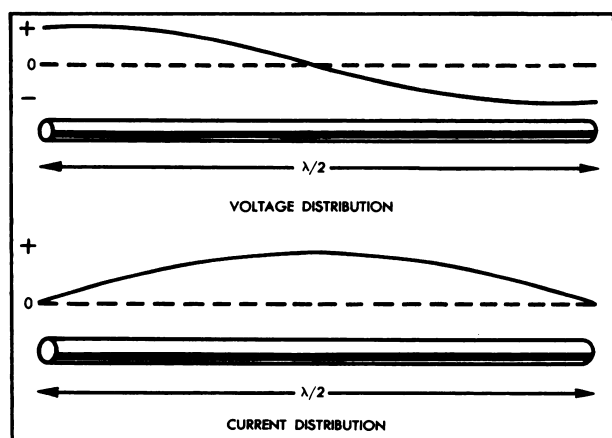


Figure 13-6. Current and Voltage Distribution in Half-Wave Antenna

An examination of the current distribution curve shows that the current standing wave reaches maximum a quarter-cycle after the voltage reaches maximum at which time the current is maximum at the center and zero at the ends. In practical applications the ends of a half-wave antenna must be insulated due to the high voltages there, and the center of the antenna must have low resistance in order to minimize the  $I^2R$  losses because of the high current there.

### RESONANCE AND DIMENSIONS

Electromagnetic waves travel through space at a speed of 300 million meters per second. The length of one cycle in space depends upon frequency and is called the *wavelength*. Mathematically, the length of an electromagnetic wave is expressed by the formula,

$$\lambda = \frac{3 \times 10^8}{f}$$

where  $\lambda$  is the wavelength in meters, and  $f$  is the frequency in cycles per second.

Due to the resistance of the wire, the movement of waves along a wire (antenna) is somewhat slower than wave movement in space. Wavelength on a wire, therefore, is slightly less than that of a wave traveling in space and is expressed by the equation,

$$l = \frac{3 \times 10^8 \times 0.94}{f} \text{ meters.}$$

Physically, an antenna is about 6% shorter than a half-wave traveling in space. (This accounts for the fact that it is necessary to multiply the wavelength in the antenna formula by 0.94.) In this manual assume that the correction in antenna length has been made whenever the length of the antenna is given in wavelengths.

### Effect of Length on Antenna Impedance

An antenna of the correct length acts as a resonant circuit and presents pure resistance to the excitation circuit. (See Figure 13-7A.) An antenna having other than the correct length displays both resistance and reactance to the excitation circuit. An antenna slightly longer than a half-wave, for example, acts as an inductive circuit. This is understandable if you think of the antenna in terms of a quarter-wave r-f line. When the antenna is excited at the center, it is

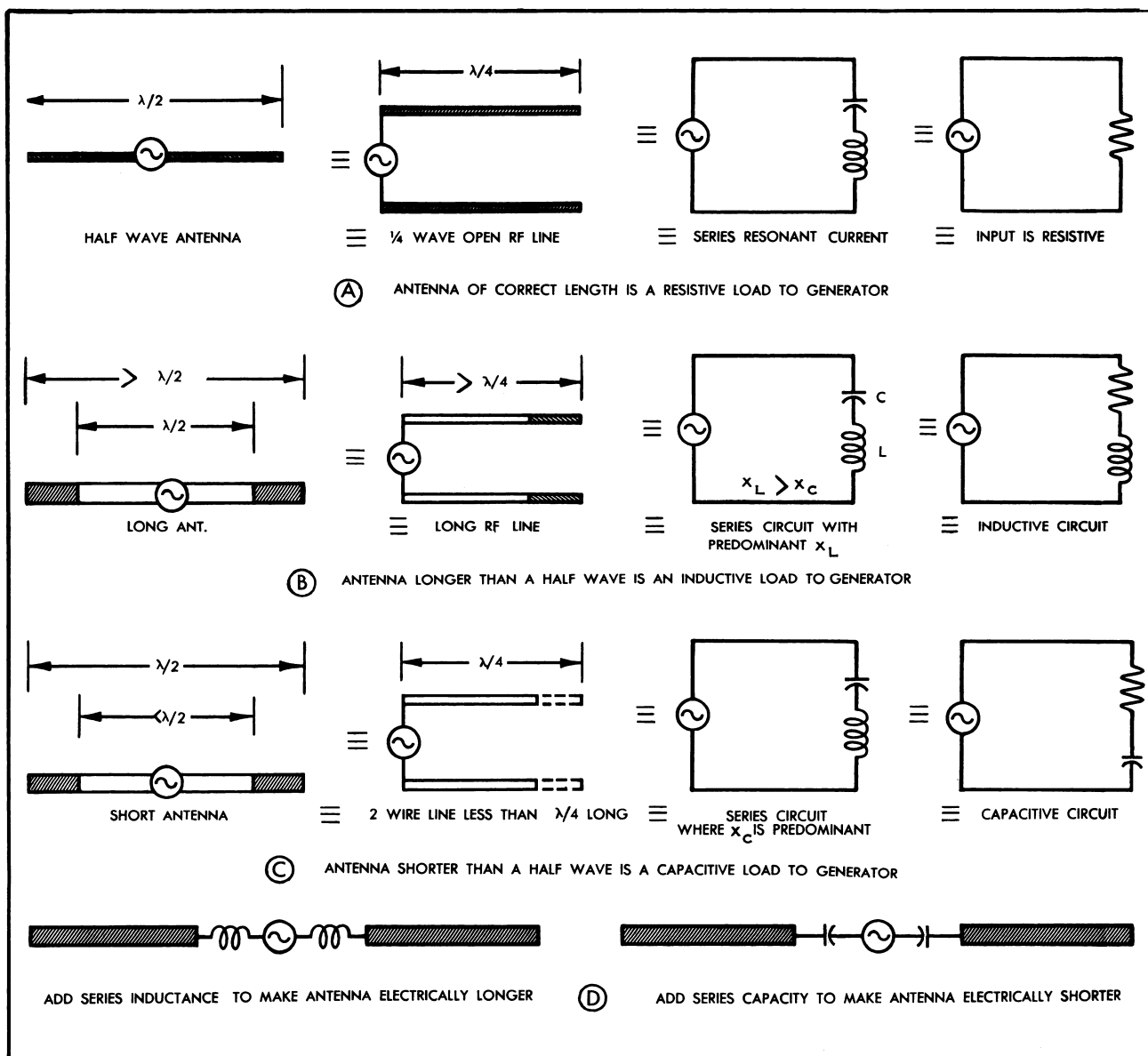


Figure 13-7. Effect of Length on Antenna Impedance

equivalent to a quarter-wave r-f line, looking at it from the generator end. Any two-wire line, longer than a quarter-wave (but less than a half-wave), is like a quarter-wave section with an open-circuited section (less than a quarter-wave long) added to it. The open section, which is capacitive, is inverted to inductance at the input terminals by the quarter-wave. In the same manner a slightly long antenna looks inductive.

According to the equivalent lumped circuit at B, the inductance is not entirely balanced by the capacitance. The remedy for correcting the

length is either cutting the conductor shorter, or tuning out the inductive reactance by adding capacitance in series.

If the antenna is physically shorter than its resonant length, the input impedance becomes capacitively reactive. The two-wire line shorter than a quarter-wavelength displays capacitance at its input terminals. (See Figure 13-7C.) The correction in this case is either to add inductance in series with the antenna to bring it back to resonance or to add physical length to the antenna.

## Increasing Antenna Length

With radar equipment a wide band of frequencies — rather than a single frequency — must be handled by an antenna. This means that an antenna which is designed for the center frequency is short for frequencies in the lower (low frequencies) sidebands, and long for frequencies in the upper (high frequencies) sidebands. The antenna must be corrected for both long and short conditions at the same time.

Note in Figure 13-8 that each side of an antenna must be less than a quarter-wave (in space) in order to present zero reactance. This agrees with the previous statement that the physical length of the half-wave element must be 6% shorter than the half-wave in space. Two curves are shown. One is for a large diameter antenna, and the other for a small diameter antenna. Note that the larger diameter displays less reactance at lengths off resonance.

Antenna diameters as great as  $\frac{1}{20}$  wavelength are found in radar equipment. A large diameter increases the capacitance of the antenna. The inductance is decreased for the same resonant frequency. Lower inductance with the same resistance lowers the  $X_L/R$  ratio, or  $Q$ , of the antenna. A lower  $Q$  causes the resonance curve to be broader and gives the antenna a more uniform response to a greater band of frequencies.

## IMPEDANCE AT INPUT TERMINALS

In practice an antenna acts as a resonant or near resonant circuit. As a series resonant circuit, its input terminals display zero impedance;

as a parallel resonant circuit, it displays infinite impedance. Further, if it is not of the correct length, it has reactance as well as resistance.

Practical antennas, like practical resonant circuits, have losses which must be replaced by the source. These losses make the input impedance of the antenna somewhat resistive. This resistance is a combination of two resistances — the resistance of the conductors themselves, which is increased by the skin effect at high frequencies, and the radiation resistance.

Radiation resistance is a fictitious resistance that dissipates the same amount of power in the form of heat as is actually dissipated as radiated energy. Because of radiation resistance, an antenna allows part of its fields to escape into space. This makes it necessary for additional current to flow into the antenna from the source of power to replace the part of the field that escaped. Therefore, the source must provide the energy lost by radiation. The energy lost is power ( $P$ ) and its ratio to the square of the current ( $I$ ) is the radiation resistance,  $R_r$ . Mathematically, it is expressed,

$$R_r = \frac{P}{I^2}$$

Radiation resistance and conductor resistance constitute the total input resistance to an antenna. It is possible to make the conductor resistance small in comparison to the radiation resistance by using large diameter, low resistance conductors. It is desirable that the radiation resistance be a major portion of the input resistance. In practice radiation ratios as high as 9:1 are obtainable. This means that an antenna can be 90% efficient as a radiator.

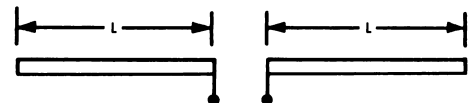
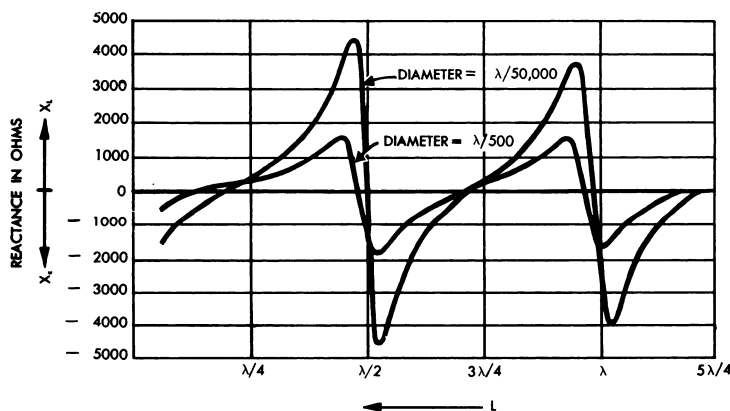


Figure 13-8. Reactance at Input of a Center-Fed Antenna of Arbitrary Length

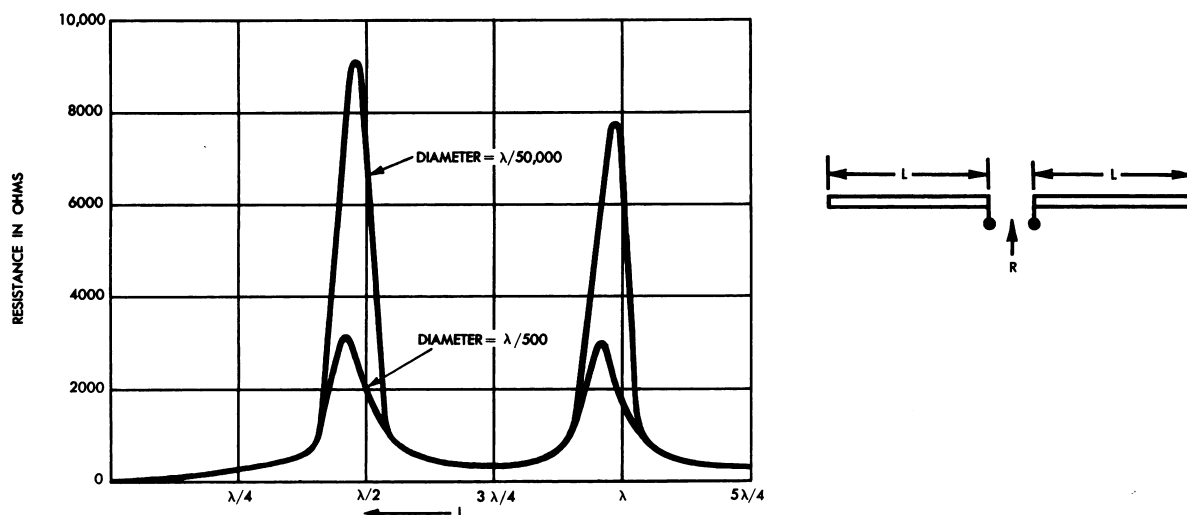


Figure 13-9. Resistance at Input of a Center-Fed Antenna of Arbitrary Length

Figure 13-9 is a graph showing the input resistance at the center of antennas of various lengths. For the half-wave dipole ( $L = \lambda/4$ ), the input is 73 ohms. The input resistance of a full-wave antenna ( $L = \lambda/2$ ) can be as high as 9000 ohms, depending on the diameter of the conductor.

For the lengths of antennas in which the reactance is zero, the input resistance is the input impedance of the antenna. However, for lengths in which the reactance is not zero, the input impedance is the vector sum of the resistance (Figure 13-9) and the reactance at the input of the antenna (Figure 13-8).

Other factors which affect the input impedance of an antenna are nearby conducting objects, such as the earth, and other antennas or the skin of the aircraft. The graphs in Figure 13-8 and Figure 13-9 apply only to a center-fed antenna which is located in free space and not close to any conductor.

## POLARIZATION OF AN ELECTROMAGNETIC WAVE

The radiated energy from an antenna is in the form of an expanding sphere. A small section of this sphere is called a *wave front*; it is perpendicular to the direction of travel of the energy. All energy on this surface is in phase. Usually all points on the wave front are at equal distances from the antenna. The farther from the antenna, the less spherical the wave appears. At a con-

siderable distance the wave front can be considered as a plane surface at right angles to the direction of propagation.

## Linear Polarization

The radiation field is made up of magnetic and electric lines of force which are always at right angles to each other. Most electromagnetic fields in space are said to be *linearly polarized*. The *direction* of polarization is the direction of the electric vector. That is, if the electric lines of force are horizontal, the wave is said to be *horizontally polarized* (Figure 13-10A), and if the E lines are vertical, the wave is said to be *vertically polarized*. As the electric field is parallel to the axis of the dipole, the antenna is in the plane of polarization. The horizontally-placed antenna in Figure 13-10 produces a horizontally-polarized wave, and the vertically-placed antenna produces a vertically-polarized wave.

For maximum absorption of energy from the electromagnetic fields, it is necessary that a dipole be located in the plane of polarization. This places the conductor at right angles to the magnetic lines of force that are moving through the antenna and parallel to the electric lines.

In general, the polarization of a wave does not change over short distances. Therefore, transmitting and receiving antennas are oriented alike, especially if short distances separate them.

Over long distances, the polarization changes. The change is usually small at low frequencies,



but at high frequencies the change is quite rapid.

With radar transmissions, a received signal is a reflected wave from some object. As the polarization of the reflected signal varies with the type of object, no set position of the receiving antenna is correct for all returning signals. Where separate antennas are used, the receiving antenna is generally polarized in the same direction as the transmitting antenna.

When the transmitting antenna is close to the ground insofar as propagation is concerned, vertically-polarized waves cause a greater signal strength along the earth's surface. On the other hand, antennas high above the ground should be horizontally polarized to get the greatest signal strength possible to the earth's surface. For this reason most airborne radar systems emit horizontally-polarized waves.

### Circular Polarization

A factor that determines the effective range of a radar set is the manner in which the r-f energy is radiated into space. The radiation of electromagnetic energy during normal weather conditions gives good target returns. However, during

a heavy rainstorm the individual raindrops act as targets and reflect the polarized electromagnetic energy, which may completely illuminate the indicator. This condition is undesirable and is overcome by radiating energy in a type of polarization that is designed to cancel returns from symmetrical targets. This type of polarization is *circular* polarization.

Imagine the electric lines of force in vertical polarization being changed from their fixed position in space and rotated through  $360^\circ$  with every cycle of r-f energy fed to the antenna. Constant amplitude electric lines of force are continuously changing their angular position as they are radiated into space. The resulting effect of this circular motion of the electric field is called circular polarization.

When a reflecting surface is hit by a cycle of r-f energy, currents are induced in the reflector. These induced currents, in turn, cause the reflecting surface to act as a transmitting antenna, radiating a cycle of energy at the same frequency as the original energy, polarized in the same plane but reversed in phase. The intensity of this transmitted (reflected) energy will be dependent

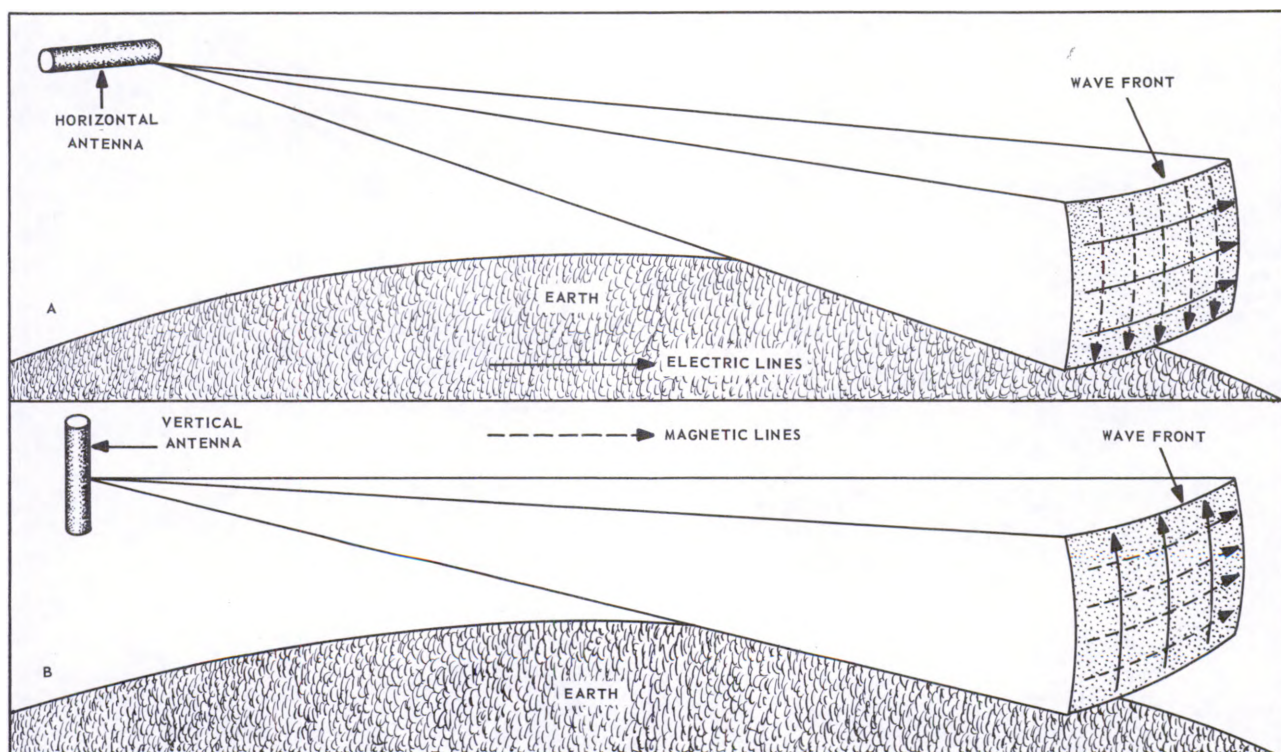


Figure 13-10. Horizontal and Vertical Polarization

on (1) the intensity of the energy that induced the current, (2) the efficiency of the reflecting surface as an antenna at that frequency, and (3) the response of the target to energy polarized in that plane.

A given reflector might be a very good radar target for vertically polarized energy and yet be a very poor radar target for horizontally polarized energy at the same frequency. Symmetrical reflectors, such as spheres, however, would respond to vertically and horizontally polarized energy equally well. In fact, they give equal response to energy polarized in any plane. When the plane of polarization changes, constantly rotating  $360^\circ$  with each cycle radiated, symmetrical targets give constant returns. On the other hand, nonsymmetrical targets, such as aircraft, do not respond equally well to the energy polarized in any plane. When the energy is polarized in one plane, the reflected energy may

be very weak. This results in returns of different amplitudes.

**POLARIZER.** The polarizer is designed to reject the circularly polarized waves reflected by raindrops and accept the energy reflected from a nonsymmetrical target (elliptically polarized returns, which are stronger in one plane than the other). Although there are several methods that will give the electromagnetic energy circular polarization, we will explain only a typical system. The piece of equipment that changes either vertical or horizontal polarization to circular polarization (Figure 13-11) is called the polarizer. The polarizer is a circular section of waveguide containing a fixed dielectric slab. The material used for the slab, *rexolite*, is known for its dielectric strength and ability to withstand the high voltages that exist across the waveguide. The polarizing feed is physically an extension of the waveguide.

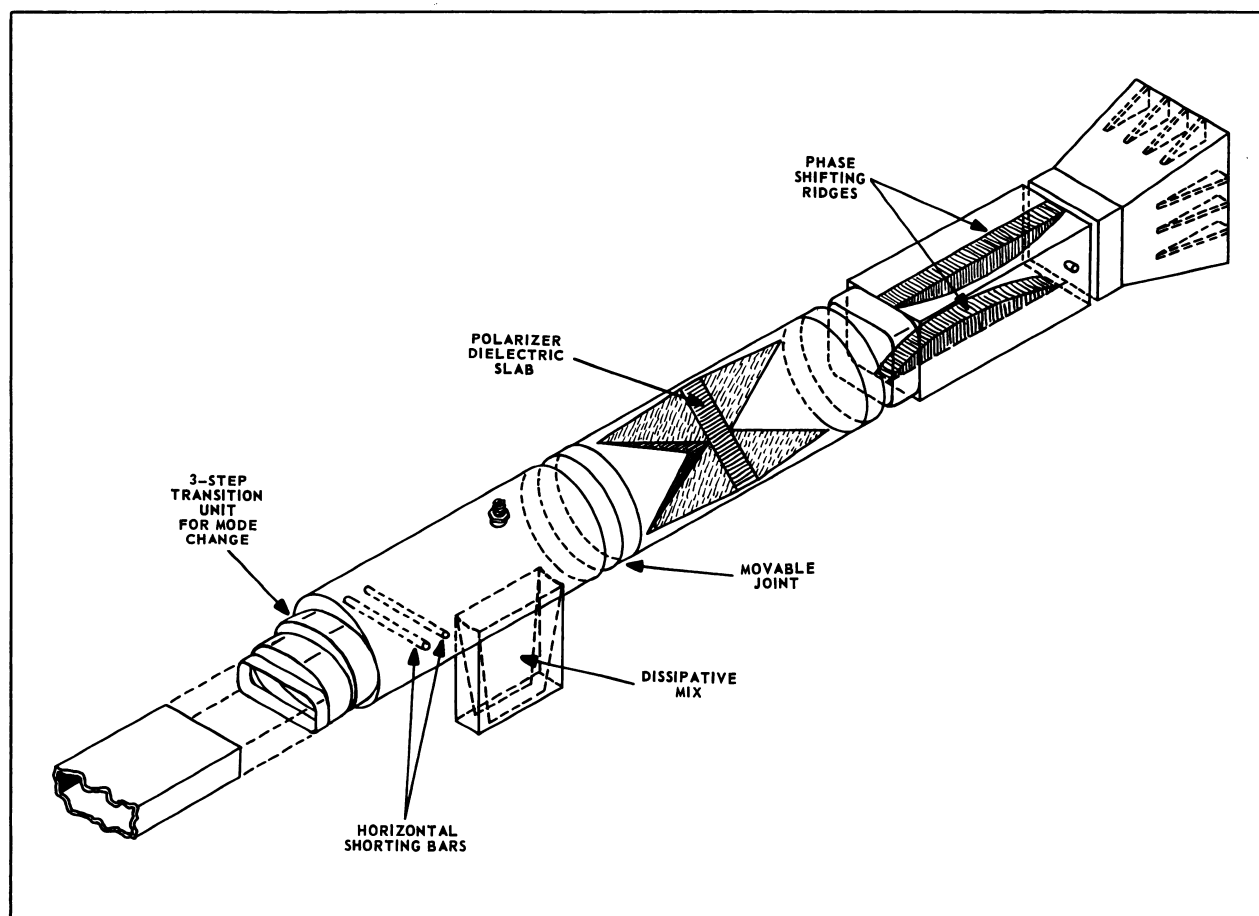


Figure 13-11. Simplified Diagram of Polarizer



A three-step change is made in connecting the rectangular waveguide to the circular waveguide, thus converting from the  $TE_{0,1}$  mode of operation to the  $TE_{1,1}$  mode. (See Chapter 11, Analogy of Waveguide Action by Electric Waves.) To see why this is true, represent the energy being transmitted through the waveguide by a vertical vector,  $V_A$ , as shown in Figure 13-12A. Now break the vector into two perpendicular components,  $V_B$  and  $V_C$ , Figure 13-12B. These two components may be assumed to represent the two sine waves shown. As long as these two sine waves are in

phase, they will combine to form the original sine wave represented by vector  $V_A$ . Furthermore, they will remain in phase as long as neither of them is intercepted by the dielectric slab. This is the condition for vertical polarization.

The three-step move makes a smooth transition and thus minimizes the reflections that come with rapid changes in the waveguide. The movable section of the guide is the polarizer, and can be rotated  $45^\circ$  to produce either vertically- or circularly-polarized r-f energy. To make the change in the dielectric material as gradual as

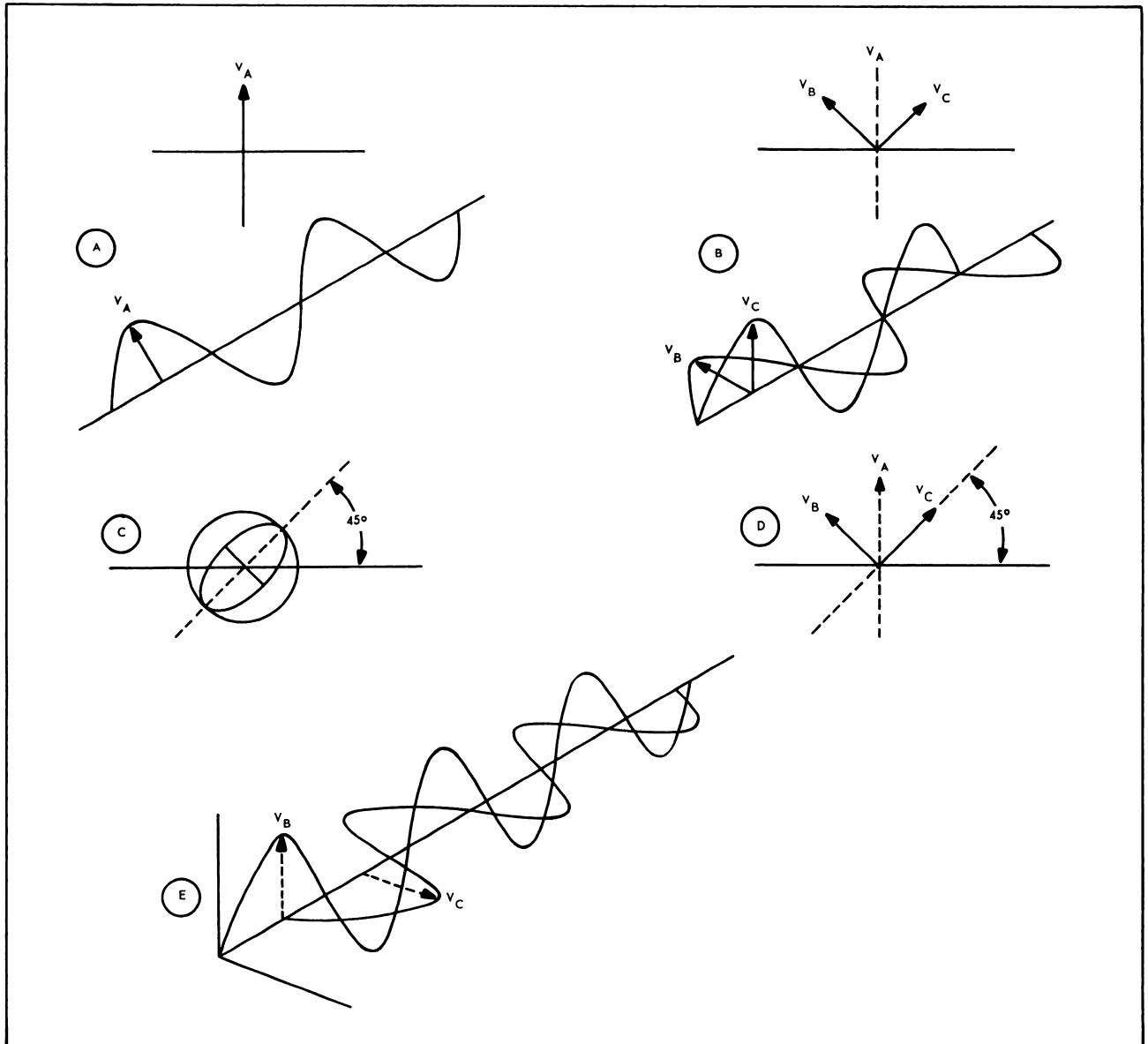


Figure 13-12. Development of Vectors and Phase Shift

possible and obtain minimum reflections, the slab is sloped at each end (Figure 13-11). The position of the slab in the waveguide determines the degree to which the energy is intercepted which, in turn, determines the degree of phase shift.

The dielectric slab is manually controlled for maximum operating efficiency. A reversible motor, through a gear reduction system, can also change the slab from its normally horizontal position. Adjustable mechanical stops ensure proper alignment of the waveguide in either the vertical or circular polarizing position. The polarized energy then passes on through a one-step circular-to-rectangular (square) waveguide transition. As the diameter of the round guide and the distance between the walls of the square waveguide are nearly equal, the transition can easily be made in one step. The square waveguide is constructed with phase-shifting ridges that prevent the circularly-polarized energy from becoming elliptical. The polarized energy passes into a feed horn that feeds the microwave energy to the reflector.

**DIELECTRIC SLAB EFFECTS.** The dielectric slab is normally horizontal, effectively perpendicular to the vertically-polarized electric field. The slab is shaped to have minimum effect on the electric field when in this position, and any small effect that the element does have will act equally on the entire cross section of the electric field. When the dielectric slab is in the horizontal position, it is perpendicular to all lines in the electric field, intercepting every line equally, so that the sine waves represented by vectors  $V_B$  and  $V_C$ , Figure 13-12, remain in phase.

On the other hand, if circular polarization is desired, one of the vectors, either  $V_B$  or  $V_C$ , must be intercepted by the dielectric slab. If the slab is rotated to an angle of  $45^\circ$  with the horizontal, as shown in Figure 13-12C, then the component vector  $V_C$  will be intercepted by the slab (Figure 13-12D). The operation of the slab is very similar to that of a phase shifter, whose action is based on the fact that the wavelength of energy in a waveguide depends upon the dielectric used within the waveguide.

In the waveguide air is the dielectric, and the wavelength of the propagated energy is shorter than it would be in free space. The slab appears to the energy vector  $V_C$  as a more dense dielectric, resulting in a shorter wavelength for the sine

wave represented by vector  $V_C$ . A shorter wavelength for  $V_C$  has the same effect in the waveguide as a phase shift for  $V_C$ . The number of degrees of shift in the sine wave represented by  $V_C$  is dependent upon the length of the dielectric slab. As a result there are two sinusoidal components which are  $90^\circ$  out of phase with each other. The two quadrature components of the electric field are indicated, showing instantaneous directions and amplitude, in Figure 13-12E.

The resultant field (vector sum) at any instant is an electric field of constant amplitude changing its direction of polarization at a uniform rate through  $360^\circ$  with every cycle of radiated r-f energy. Figure 13-13A illustrates the resultant waveform with the arrows depicting the electric field. Figure 13-13B shows the direction of the E field at a point in space at various instants of time over a complete cycle.

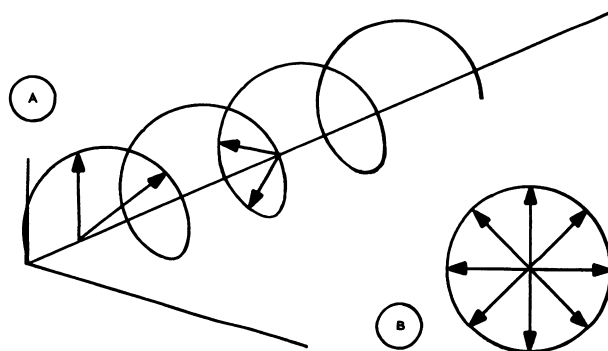


Figure 13-13. Wave Depicting Circular Polarization and Vectors

When the circularly-polarized energy hits a symmetrical target (a raindrop) and returns, the two quadrature components remain equal in amplitude, both having been changed the same amount. This is because the raindrop is essentially a sphere and therefore has a symmetrical surface. As the reflection from the raindrop returns to the polarizer, part of the energy (vector  $V_C$ ) is again intercepted by the dielectric slab. Once again the sine wave represented by  $V_C$  is shifted  $90^\circ$ . Since it had already been shifted  $90^\circ$  during the transmit cycle, this wave is now  $180^\circ$  out of phase with the sine wave vector as represented by  $V_C$  in Figure 13-12D. Figure 13-14A represents the return vectors and Figure 13-14B the sine waves after returning

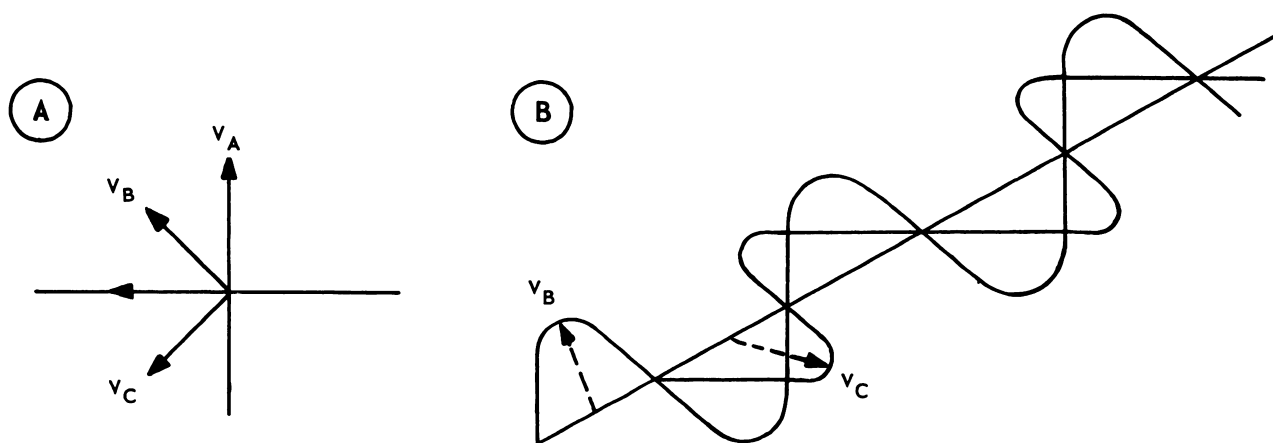


Figure 13-14. Return Phase Shift for  $V_0$

through the polarizer. Note that the resultant energy is now horizontally polarized. Horizontal shorting bars (Figure 13-11) cancel these returns, and the power is dissipated in the filter load (Figure 13-11).

When the circularly-polarized energy strikes a target that is not spherical, the results are different. Perhaps it will fail to reflect all of the energy represented by the vector  $V_C$ . In this case the sine wave resulting from vector  $V_C$  will have a different amplitude from that of the sine wave resulting from vector  $V_A$ . The resultant will have a vertical component which, as the energy passes through the polarizer, will *not* be shorted by the horizontal shorting bars but instead will pass on to the receiver. This is illustrated in Figure 13-

15A with  $V_x$  being the vertical component. This is called *elliptical polarization*. Where the magnitude of the electric vector changes over the cycle, the wave is said to be *elliptically polarized*. Figure 13-15B illustrates the electric field at various instants of time over a complete cycle. Figure 13-15C illustrates the condition where the vertical field is maximum at  $45^\circ$ .

Tests have proven that returns from dense rain clouds can be reduced 30 db by using circular polarization. While returns from the nonsymmetrical targets are weaker, which is a radar range factor when circular polarization is used, the elimination of most interference from heavy rain storms has proved the positive value of the circular polarization feature in search radars.

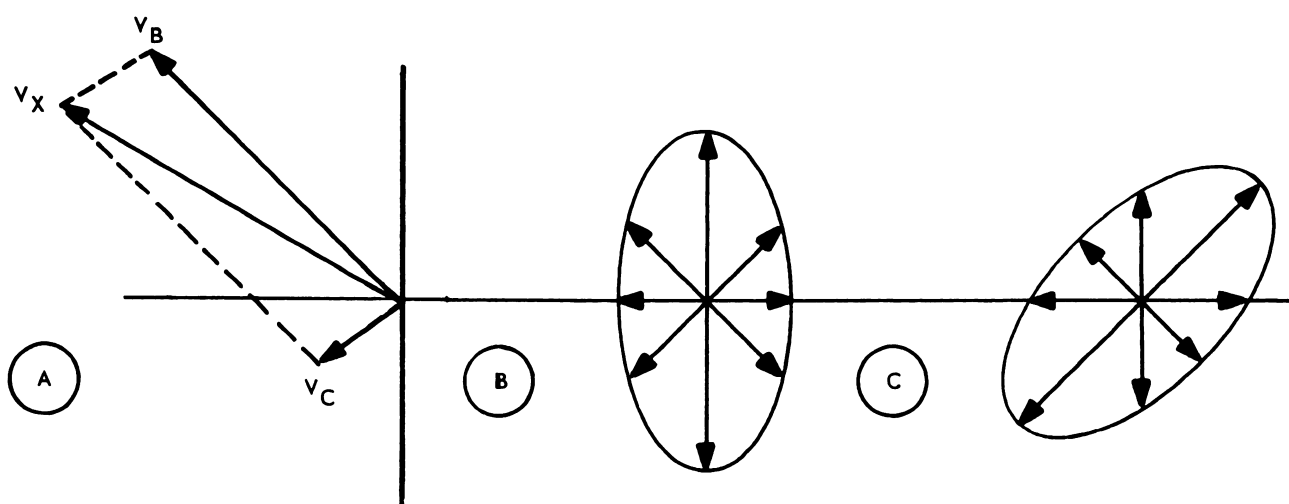


Figure 13-15. Vertical Component of Return Wave Elliptically Polarized

## DIRECTIONAL CHARACTERISTICS

If you measured the radiation from a single point in space, you would find that this point radiates equally well in all directions. Since the strength of the radiated energy varies inversely as the distance, if all the points where the energy was of the same strength were plotted, the points would form a sphere, with the radiating point at the center.

If radiation occurs from more than one point in space, the radiated signal from each point will add vectorially to produce the total radiation strength.

Note the two points in space in Figure 13-16. Each point is radiating a similar field in all directions. Concentric circles are shown around each point. The circles represent positive and negative peaks of the radiation field. The solid lines are positive peaks. For simplicity the points are placed a half-wave apart. The picture is stopped just as each point is ready to generate a negative peak.

Note the peaks as they cross the vertical line  $YY'$ . Adding the two fields there doubles the field strength. Next observe the peaks along the line  $XX'$ . Wherever a positive peak from one point appears, a negative peak from the other also appears. This means that cancellation occurs along the line  $XX'$ .

The radiation from a pair of points is directional. Most of the energy is detected along  $YY'$ , while little or no energy is detected along  $XX'$ . The strength at angles between  $X$  and  $Y$  varies as the cosine of the angle.

When you apply these facts to a half-wave antenna, you can look at the conductor as a series

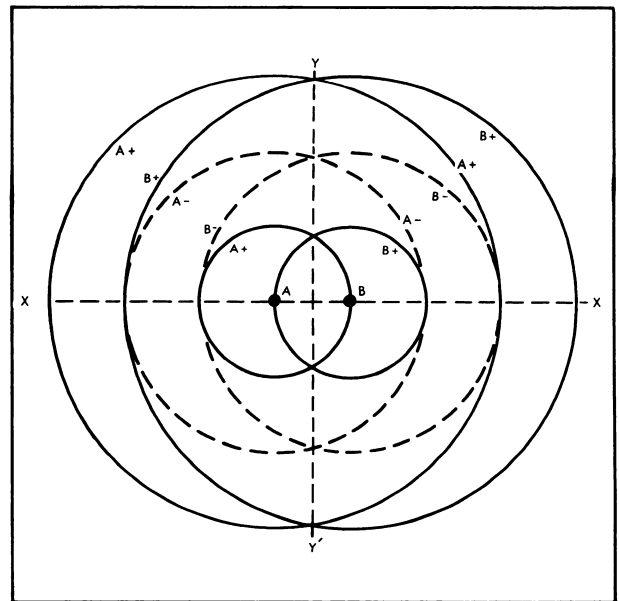


Figure 13-16. Vector Addition of Electromagnetic Energy from Points A & B

of points arranged in a straight line. The radiation from points equally spaced from the center will add in directions perpendicular to a line through them, but will cancel along the line through them. A plot of all points of equal strength would produce a figure which is doughnut-shaped. This three-dimensional figure is shown in Figure 13-17. Note that a cross section resembles two ellipses which are adjacent to one another.

The cross-sectional view is called a *polar diagram*. This sort of diagram is standard for delineating antenna performance on paper.

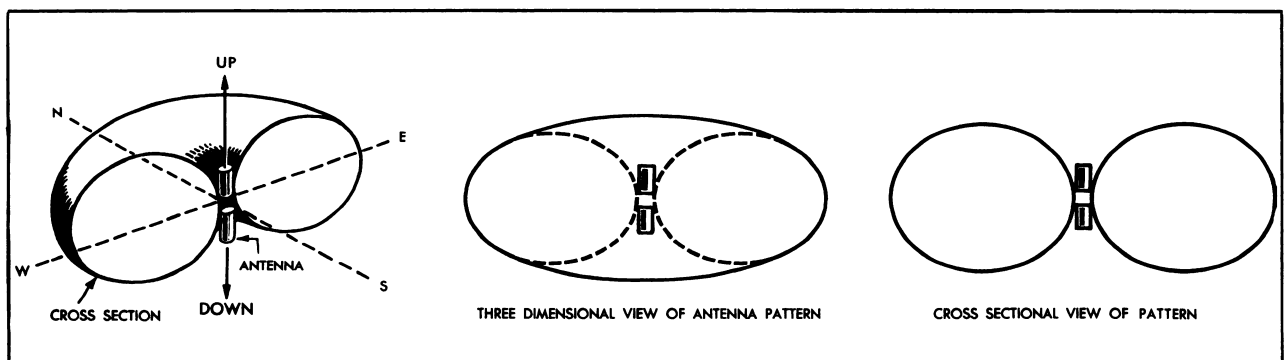
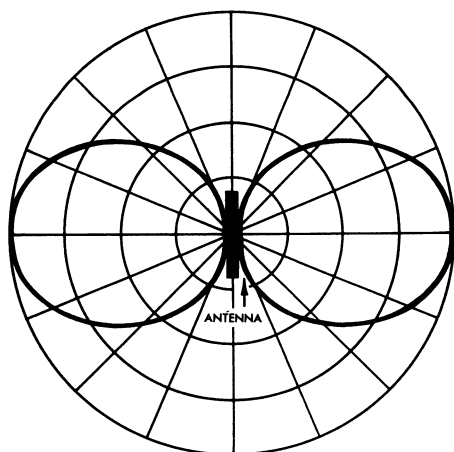


Figure 13-17. Cross-Sectional View of Vector Addition Figure

## Constructing Polar Diagram

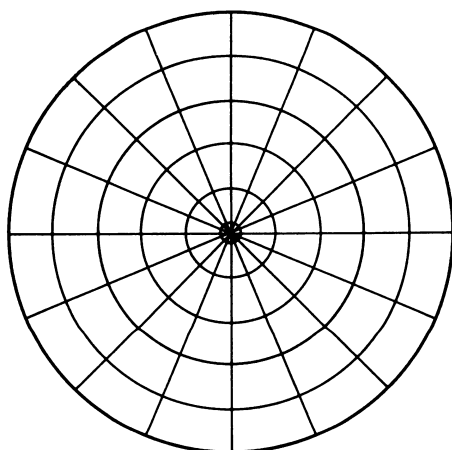
There are two methods of constructing a polar diagram. One method consists of measuring the field strength at the same distance in all directions from the antenna, and plotting the angle against field strength on polar coordinate paper. The other method consists of finding the points where the field strength is the same, and plotting the angle against distance on polar coordinate paper. Both types of measurement produce identical polar diagrams.

Note the polar diagrams for the dipole in Figure 13-18. Diagram A holds good for any plane containing the antenna. For the plane perpen-



POLAR DIAGRAM FOR HALF-WAVE ANTENNA IN ANY PLANE CONTAINING THE ANTENNA

A



POLAR DIAGRAM FOR HALF-WAVE ANTENNA IN PLANE PERPENDICULAR TO ANTENNA

B

Figure 13-18. Polar Diagram of Dipole

dicular to the half-wave antenna, the antenna forms the single point which was derived previously. The polar diagram for this point becomes a circle with the point at the center, as in diagram B. Thus, the simple dipole is bi-directional in any plane containing the antenna, but nondirectional in the plane perpendicular to the antenna.

## Field Strength

The strength of a radiation field is called *field strength* and is measured in units called *volts-per-meter*. One volt-per-meter is defined as the field strength that will induce one volt in a conductor one meter long. As field strengths of a volt-per-meter are seldom encountered, weaker fields are measured in *microvolts-per-meter* or *millivolts-per-meter*.

## IMAGE ANTENNAS

The preceding discussion dealt with antennas that are isolated from any conductor. However, in all practical cases, there are usually conductors somewhere near the antenna. The effect of the conductors is often undesirable but can sometimes provide advantages.

## Theory

The effect of a nearby conductor is illustrated diagrammatically in Figure 13-19A. A real antenna is shown perpendicular to a horizontal per-

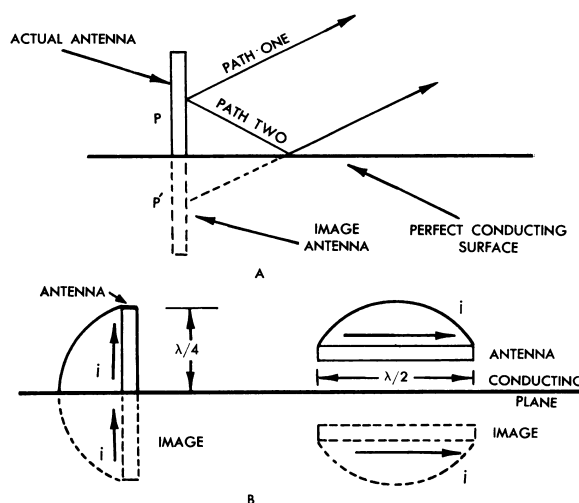


Figure 13-19. Image Antennas

fectly conducting plane. When a field is radiated, energy will be sent out in all directions. There are two possible paths for the field to take. Path one starts out directly toward a distant point. The energy in path two starts out toward the conducting plane. Upon striking the perfect conductor, it is reflected in reverse phase. (This is analogous to an incident wave being reflected in reverse phase from the shorted end of an r-f line.)

As with light waves, the angle of reflection is the same as the angle of incidence. The reflected wave also proceeds toward some point in space. The sum of the two waves makes up the total wave at any point in space. So far as the action in the conducting plane is concerned, it can be replaced by another antenna which is a mirror image of the actual antenna. The reflected wave can be assumed to have originated at point P' on the image antenna.

In general, current in vertical image antennas flows in the same direction as in actual antennas, while in horizontal antennas the current flows in opposite directions. Note in Figure 13-19B that the combination of the real and image antennas for a vertical quarter-wave makes a dipole.

### Marconi Antenna

A quarter-wave grounded antenna is a common type of grounded antenna. This type is called a *Marconi* antenna, as contrasted with the half-wave (ungrounded dipole) antenna, which is called a *Hertz* antenna. Note the amplitude of the standing waves of current and voltage on the Marconi antenna shown in Figure 13-20A. Note also the similarity to the half-wave element when the image is included.

In the Marconi antenna the vertical field strength pattern (polar diagram) shown in Figure 13-20B is the same as that of the half-wave element, except that the conducting plane cuts it off at the center. The image is only effective above the plane because no energy penetrates the conducting plane.

In the horizontal polar diagram in Figure 13-20C the vertical field strength pattern can be rotated with the antenna as the axis to form the horizontal polar diagram of a Marconi. It is nondirectional in a plane perpendicular to the length of the antenna.

The input impedance to the Marconi is approximately 37 ohms when the antenna is fed at its base as illustrated in Figure 13-20A. In

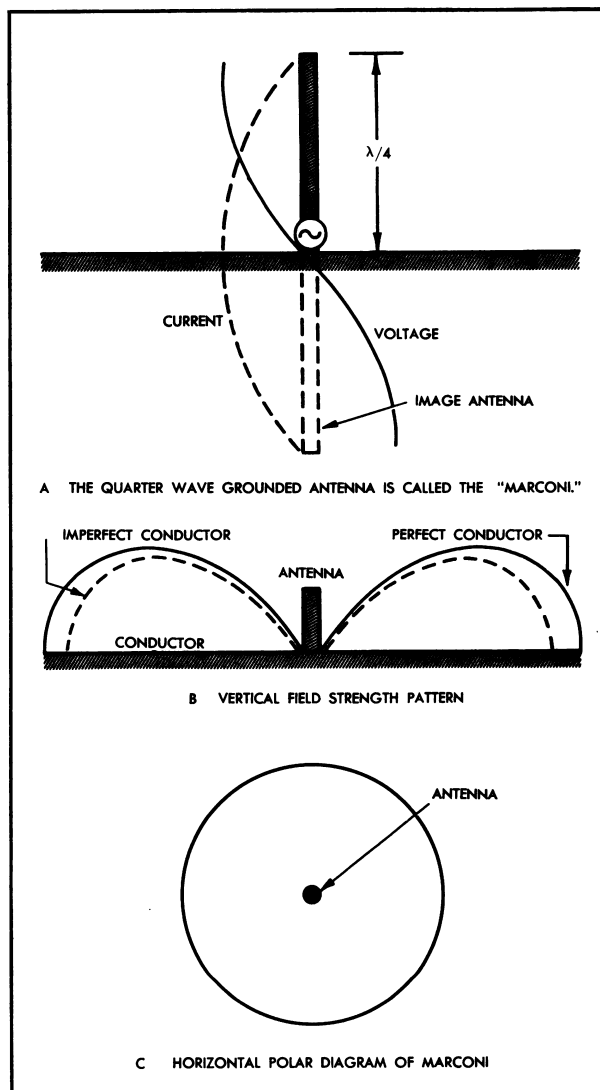


Figure 13-20. Marconi Antenna

addition a quarter-wave Marconi is resonant and displays zero reactance just like a half-wave antenna.

This discussion has assumed that the conducting plane is a perfect conductor. If it is not a perfect conductor, as is the usual case, some of the conditions just discussed must be altered. The principal alteration is the reduction in size of the polar diagram. This results in decreased signal strength from the antenna (Figure 13-20B). The conducting plane is usually the skin of the aircraft where airborne equipment is concerned; with ground equipment it is the earth's surface, or an artificial ground called a *counterpoise*.



Other types of conductors that might be near an antenna are the aircraft tail pieces, the aircraft wing, or steel antenna towers. As the radiation field passes any conductor, currents are induced in the conductors. These currents vary at the same radio frequency and make the conductor itself a radiator. In other words, when the conductor is excited it reradiates, creating a radiation field of its own. This is the way in which radar "echoes" are produced.

When the radiation field of the conductor is strong, it will change the radiation pattern of the original antenna in two ways. First,  $I^2R$  losses in the conductor absorb power from the radiation field. This reduces the field strength in areas beyond the conductor. Second, the re-radiated field combines with the antenna radiation field producing random phase relationships. If they occur in phase, a strong resultant field will be present in the area. If they are out of phase, the field strength will be low. The presence of conductors is undesirable unless they are deliberately placed near the antenna to distort the field strength pattern. This is discussed later under *arrays*.

## INTRODUCING ENERGY TO THE ANTENNA

Energy may be fed to an antenna in a variety of ways; the most common are shown in Figure 13-21. When the excitation energy is introduced to the antenna at the point of high circulating currents, the antenna must be *current-fed*. When the energy is introduced at the point of maximum voltage, the antenna must be *voltage-fed*.

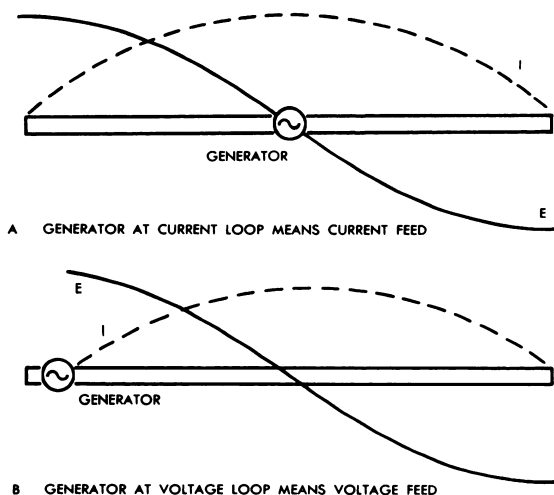


Figure 13-21. Current and Voltage Feed

It is seldom possible to connect a generator directly to an antenna. Instead, it is necessary to use r-f lines to carry the energy from the generator to the antenna. The r-f lines which carry the excitation energy may be resonant lines, nonresonant lines, or a combination of both.

## Feeding Antennas with Resonant Lines

In a voltage-fed half-wave antenna that is excited by a resonant r-f line, as shown in Figure 13-22A, the end of the antenna is connected to one side of the line. Voltage changes at that point excite the antenna into oscillation. The impedance at the end of the antenna is high, since there the voltage is high and the current is low. The r-f line shown is a half-wavelength two-wire line. The impedance of such a line is usually low so that the high impedance at which it is terminated will set up standing waves on it. In addition, it is always a multiple of half-wavelengths electrically in order that it will act as a resonant circuit. This makes the input to the line also a high impedance. A parallel resonant circuit is used to develop a high voltage across this high input impedance. Small irregularities in line lengths can be compensated for by tuning the parallel resonant circuit at the input.

In the current-fed antenna with a resonant line shown in Figure 13-22B, the r-f line is connected to the center of the antenna. This antenna has a very low impedance at the center and, like the voltage-fed antenna, has standing waves on it. It also is a multiple of half-wavelengths and is a resonant circuit. Making it exactly a half-wavelength causes the impedance at the sending (generator) end to be low. A series resonant circuit is used to develop the high current necessary to excite the line. Adjusting the capacitors at the input compensates for irregularities in line and in antenna length.

Although these examples of feeding antennas are simple ones, the principles described apply to antennas and to lines of any length — providing both are of resonant length.

The line which is connected to the antenna can be a two-wire or a coaxial line. In radar equipment the coaxial line is preferred.

One advantage of the resonant line is that it makes matching devices unnecessary. In addition, it makes it possible to compensate for incorrect antenna lengths by transmitter tuning. Its disadvantages are increased  $I^2R$  losses from

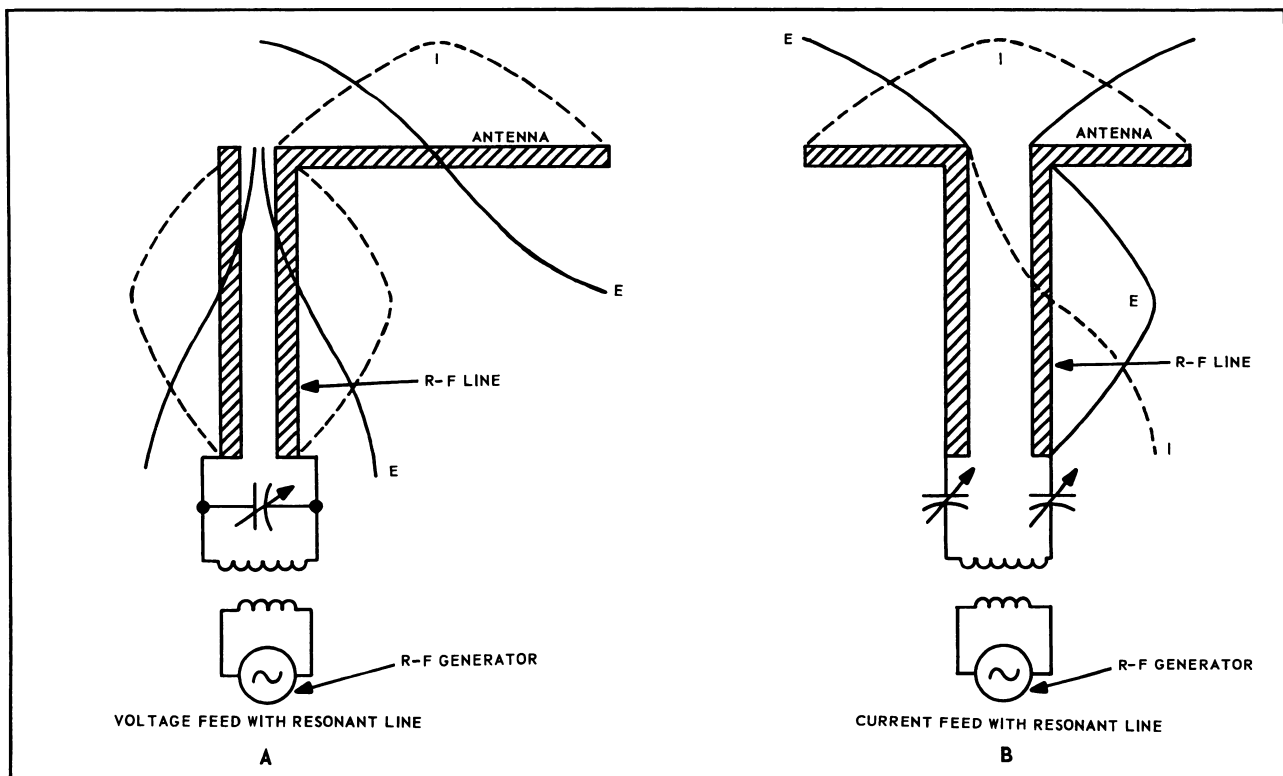


Figure 13-22. Feeding Antennas with Resonant Lines

high standing waves of current on the resonant feed lines, increased possibility of arc-over between lines due to standing waves of voltage, very critical length, and production of radiation fields by the two-wire line because of the standing waves on it. Radiation fields are undesirable because they distort the normal radiation pattern of the antenna.

#### Feeding Antennas with Nonresonant Lines

Figure 13-23 shows the excitation of a half-wave antenna by nonresonant lines. As the input to the center of the antenna at A is 73 ohms and as a coaxial cable with the highest Q has a characteristic impedance of 73 ohms, the most common method of feeding this antenna is through a coaxial cable connected directly to the center of the antenna. This method of connection produces no standing waves on the line (coaxial cable) when the generator is matched to the line. Coupling to the generator is usually made, as you see at A, through a simple untuned transformer secondary. Another means of

efficient transfer of energy to the antenna can be accomplished by a twisted pair line as shown at B. In a twisted pair the impedance is about 70 ohms. It, like the coaxial cable, provides a good impedance match.

When a line does not match the impedance of the antenna, it is necessary to use special impedance matching devices. Any of the impedance matching r-f lines discussed in Chapter 10 are adequate for this purpose. An example of a type of impedance matching device is the delta match shown at C. It gets its name because the wires form the Greek letter Delta ( $\Delta$ ) at the antenna.

The delta match is used to match a line having more than 73 ohms impedance to the center of an antenna. The line in the example shown is a 600-ohm line. The connections from the line are spread apart at the antenna until the impedance between connections is equal to 600 ohms.

As the connections are moved away from the center in the delta match, the voltage becomes

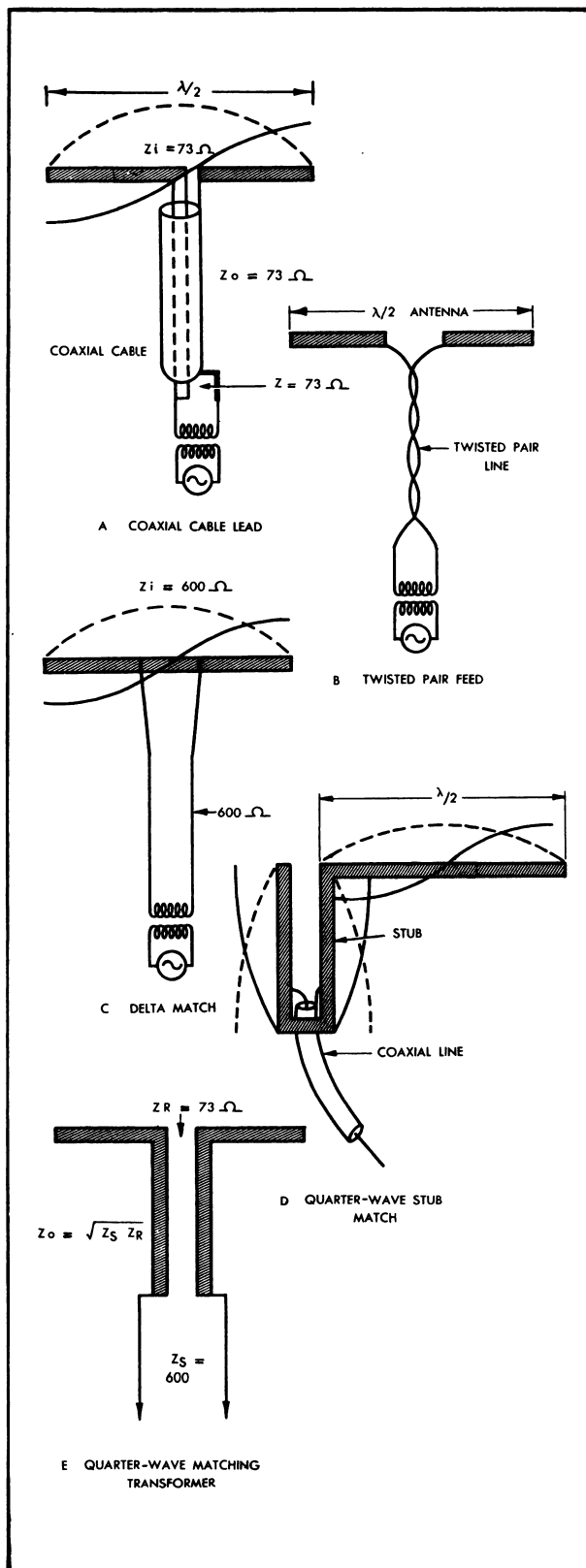


Figure 13-23. Feeding Antennas with Nonresonant Lines

higher and the current becomes lower. The ratio between these, known as the E/I ratio, increases from a very low value at the center to several thousand ohms at the ends. The two connections are set at the point where this ratio is equal to 600 ohms. This is the same as matching various impedances along a shorted quarter-wave section of line.

Another method of matching impedance is the quarter-wave stub match shown in Figure 13-23D. In the quarter-wave stub match the high impedance at the end of the antenna matches the open end of the stub. The impedance on the stub varies from zero at the short circuit to several thousand ohms at the open end. This makes it possible to connect a 70-ohm coaxial line a short distance from the shorted end at the 70-ohm point. It is possible to match almost any impedance along the length of the stub.

Still another impedance matching device is the quarter-wave transformer (Figure 13-23E). This device is used for matching the low impedance at the antenna to a line of higher impedance. In the example shown, the characteristic impedance,  $Z_o$ , of the quarter-wave section is 210 ohms. With this matching device standing waves will exist on the antenna and the quarter-wave section but not on the 600-ohm line itself.

Nonresonant lines are characterized by small radiation, low voltage at all points for any given power, and noncritical length. In addition they are preferred for the transfer of power over distances greater than one or two wavelengths.

### Exciting Antennas by Radiation Fields

The antenna may be excited either by a direct connection or by radiation (induction fields) from nearby antennas. A half-wave antenna driven by a directly connected generator is shown in Figure 13-24. The flux from the driven antenna expands and cuts the other element and induces in it a current that varies at the exciting frequency of the driven antennas and which, in turn, causes additional radiation fields. When the antenna elements are properly designed, the radiation field resulting from the induced current will be nearly as strong as the original fields.

In addition to being excited by a nearby antenna, an antenna may also be excited by radiation fields from a distant transmitting antenna. This takes place in any receiving antenna. In a

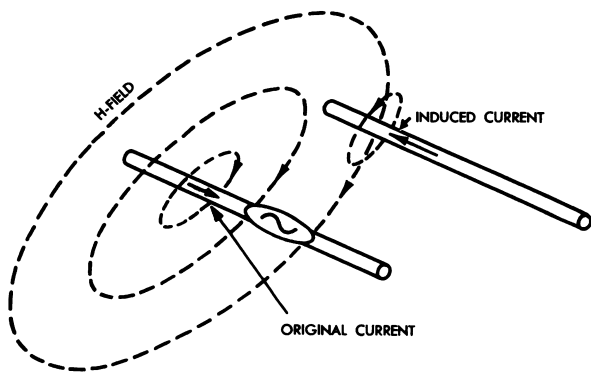


Figure 13-24. Exciting Antenna by Radiation Fields from Nearby Antenna

receiving antenna the current that is conducted to the receiver is the result of current induced by radiation fields generated by the transmitting antenna.

### ANTENNA ARRAYS

There are uses for antennas where it is desirable that the radiation be absolutely nondirectional, and other uses where the antenna must

be very sharply directional. In the latter case the antenna system usually consists of two or more simple half-wave elements so spaced that the fields from the elements add in some directions and cancel in others. The set of antenna elements is called an *antenna array*. There are two types of antenna arrays — *driven* and *parasitic* arrays.

#### Driven Arrays

The driven array consists of two or more elements with all elements connected to the generator. They may be divided into three basic types — the *broadside array*, the *end-fire array*, and the *colinear array*.

**BROADSIDE ARRAY.** When two or more half-wave elements placed one half-wave apart and parallel to each other are excited in phase, Figure 13-25A, most of the radiation is broadside to the plane of the elements. The array is termed a *broadside array*.

While the radiation around each element is nondirectional (looking at the antennas end on), these fields combine to form a directional field. When the wavefront from antenna 1 reaches an-

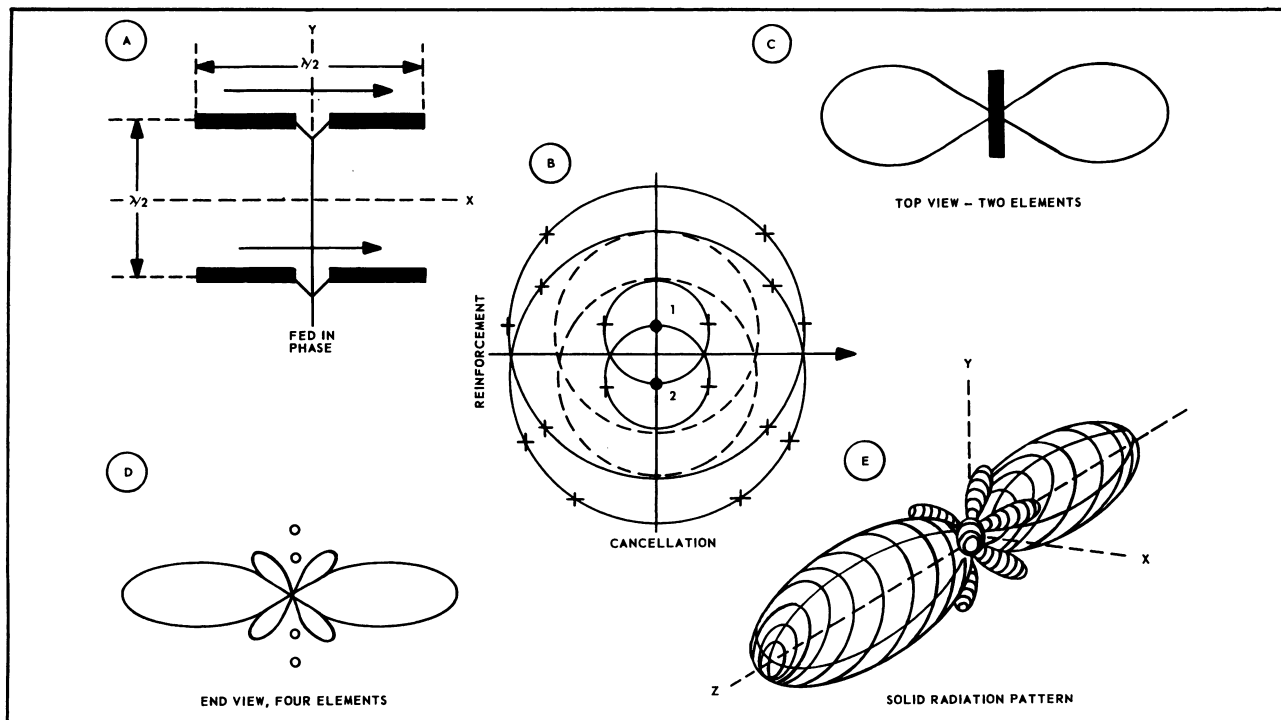


Figure 13-25. Broadside Array

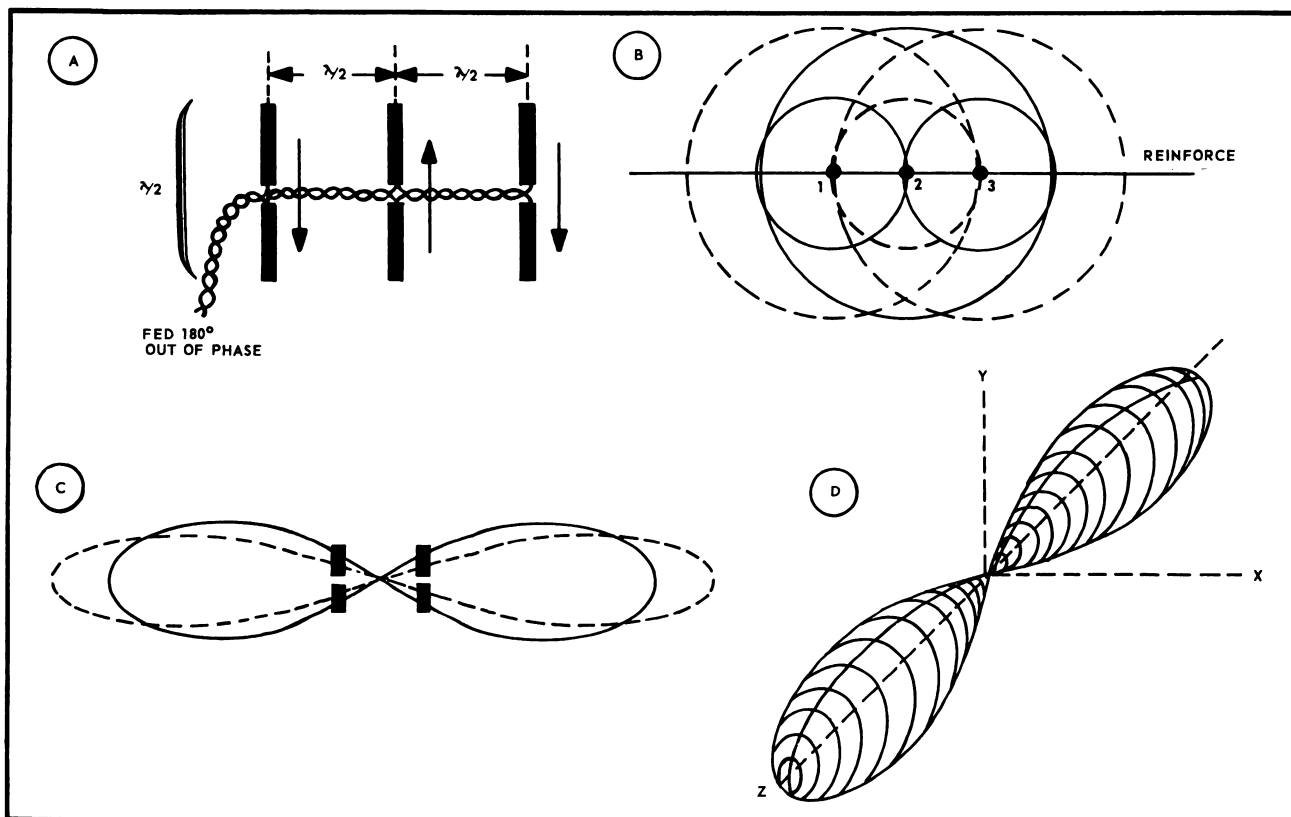


Figure 13-26. End-Fire Array

tenna 2, it is  $180^\circ$  out of phase with the energy now leaving antenna 2. As shown in Figure 13-25B, these fields cancel in the plane containing the antennas. However, in the plane at right angles to the plane of the antennas and midway between them the fields reinforce. The gain in the broadside direction, or the direction perpendicular to the plane of the antennas (each antenna is also parallel to this plane), is twice the gain of the individual antenna. Figure 13-25C is a view of the radiation pattern looking down on the antennas, arranged one above the other. As more elements are added, the pattern becomes more directional and side lobes appear (Figure 13-25D and Figure 13-25E).

**END-FIRE ARRAY.** If half-wave antenna elements spaced a half-wavelength apart and parallel are fed  $180^\circ$  out of phase (Figure 13-26A), the radiation field is strongest at the ends of the row of dipoles. This array is known as an *end-fire* array.

Figure 13-26B shows the wavefronts traveling out from each antenna, looking at the ends of the antennas. Since the antennas are a half-wave-

length apart and each is excited  $180^\circ$  out of phase with the adjacent antennas, when the wavefront from one antenna reaches the next antenna, it is in phase with the wave just leaving that antenna. Thus the radiation fields reinforce each other along the axis of the antennas.

From above the radiation pattern appears as shown in Figure 13-26C. The effect of additional elements is shown in Figure 13-26D.

**CARDIOID ARRAY.** If two dipoles are spaced a quarter-wavelength apart and are excited one-fourth cycle ( $90^\circ$ ) out of phase, a unidirectional pattern results. In Figure 13-27A the excitation of antenna *a* leads that of antenna *b* by  $90^\circ$ . When the wave from *a* reaches antenna *b*, the wave just leaving *b* will be in phase with it, due to the  $90^\circ$  lag and  $\lambda/4$  the first wave has traveled. These will reinforce each other. However, when the wave from antenna *b* reaches antenna *a*, it is  $180^\circ$  out of phase with the wave now leaving antenna *a*, due to the  $90^\circ$  lag in excitation and the quarter-wavelength it has traveled. Waves traveling toward antenna *a*, then, are  $180^\circ$  out

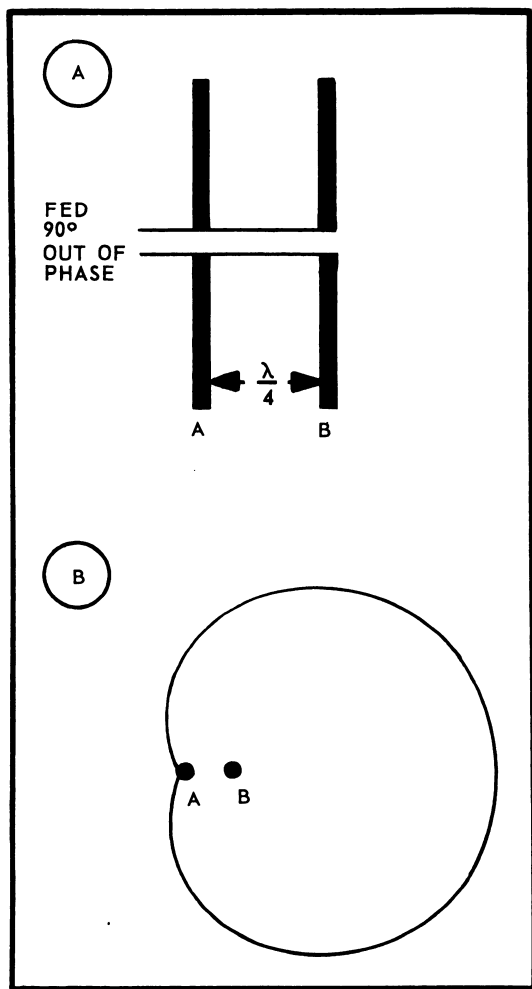


Figure 13-27. Cardioid Array

of phase and cancel out radiation in that direction. The radiation pattern appears to be heart-shaped (Figure 13-27B) so this array is often termed a *cardioid* array. The pattern is unidirectional in the direction of the antenna fed by the *lagging* excitation.

**COLINEAR ARRAY.** If two half-wave elements are placed end to end as shown in Figure 13-28A and excited in phase, they form a *colinear* array. Looking at the end of the array the pattern is circular, just as is that of a single dipole. The effect of the colinear array is to squeeze the doughnut pattern into a flat disc (Figure 13-28B). As more elements are added, the disc of radiation becomes flatter and small secondary lobes appear. The radiation pattern of a four-element colinear array is shown in Figure 13-28C.

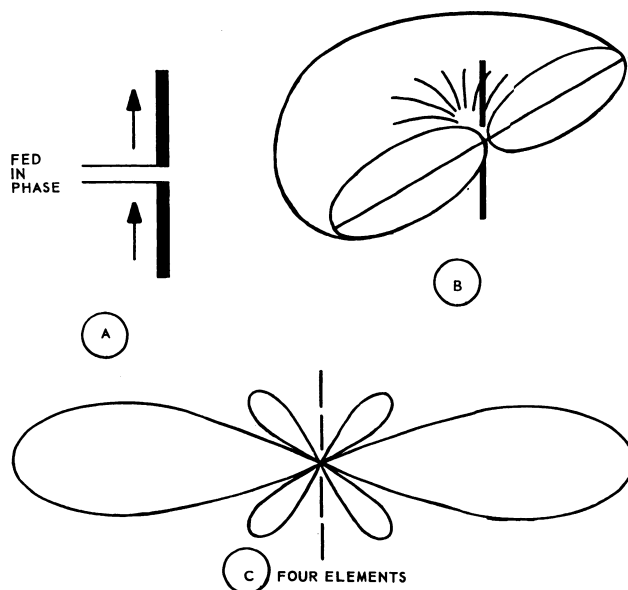


Figure 13-28. Colinear Array

#### Interconnection of Elements in Arrays

There are many ways to connect the elements in an antenna array to obtain the required phase of excitation. The most convenient method of changing the phase from one element to another element a half-wave distant is with an r-f line. In the broadside array shown in Figure 13-29A one element is excited directly from the transmitter. This element is connected to the other element by a half-wavelength of r-f line. The phase of the voltage along the r-f line is shifted  $180^\circ$  for each half-wavelength. The leads to the second driven element are reversed. This causes another  $180^\circ$  phase shift. The second element is then excited in the same phase as the first.

It is possible to connect any number of elements in this manner. In addition, increasing the number of elements makes the array more directive.

In the end-fire array in Figure 13-29B the elements are also interconnected with r-f lines. But instead of reversing the leads to the elements, they are connected directly to take advantage of the phase shift. With half-wave spacing the interconnecting line will be a half-wave long and will provide a phase shift of  $180^\circ$ . If quarter-wave spacing is used, the interconnecting line will be a quarter-wave long. This arrangement produces a  $90^\circ$  phase shift.



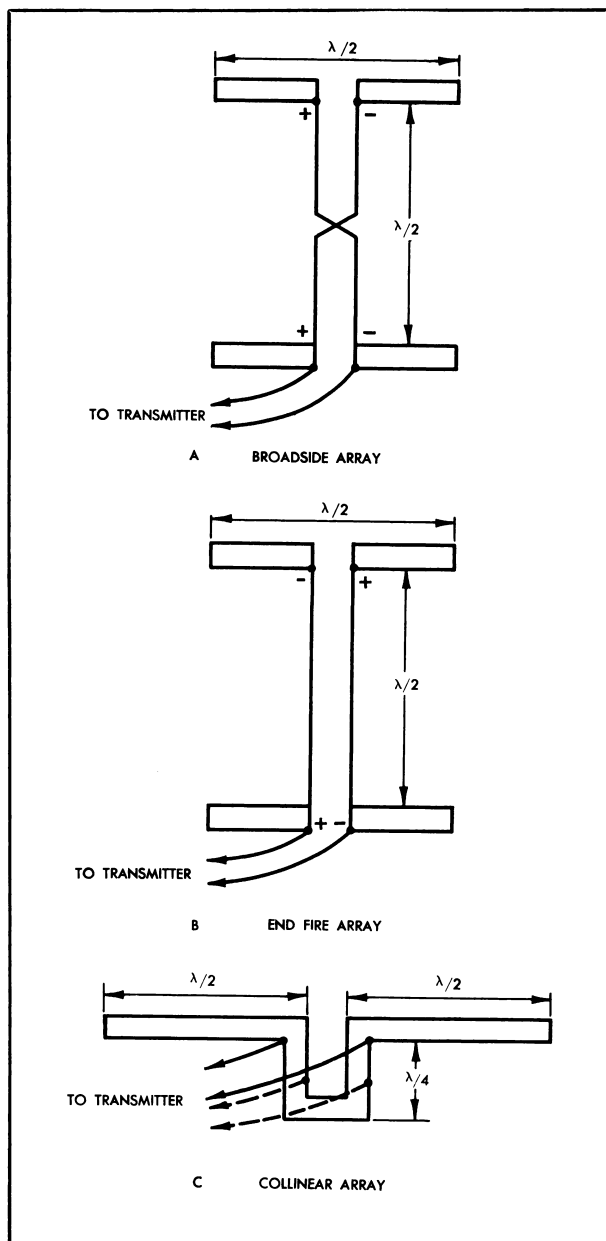


Figure 13-29. Connecting Half-Wave Elements to Obtain Proper Phase Relationship

The usual method for providing the correct phase in a collinear array is shown in Figure 13-29C. Since current direction changes for each half-wavelength of an antenna, it is not possible to connect the half-wave sections together directly. Instead, the half-wave section that carries the current in the wrong direction is folded to form a quarter-wave section of r-f line. This brings together the ends of the sections in which current flows in the same direction.

In terms of voltage, it is necessary that the two antenna sections have voltages that are of opposite polarity at the open end. Further, note that the line to the transmitter is connected at a point of very high impedance. This requires the use of a resonant line. If a nonresonant line is to be used, it must be connected as shown by the dotted lines. In Figure 13-29C the connection near the closed end of the quarter-wave section is at a point of low impedance. Another way of correctly exciting a collinear array is by a waveguide. This method is described later.

### Parasitic Array

A parasitic array is an antenna system that consists of two or more elements, one or more of which are excited by the induction and radiation fields produced by the driven element or elements. With such arrays it is possible to obtain highly directional patterns without changing the length of the antenna.

The parasitic elements are placed parallel to the driven half-wave antenna, the spacing depending upon the effect desired. Because of currents set up in the parasitic elements by the induced emf, they, too, radiate energy. At some points the energy reradiated by the parasitic element will be in phase with that radiated directly by the driven element; at other points the fields will be  $180^\circ$  out of phase, and there will be cancellation. The unidirectional pattern can be found by plotting the signal strength measured in all directions about the array. When the radiation pattern is strongest in the direction of the driven element, the parasitic element is called a *reflector*. When the maximum radiation is in the direction of the parasitic element, this element is called a *director*.

The phase difference between the current in the driven element and that in the parasitic element depends upon the distance between the driven element and the parasitic element and the relative length of the parasitic element. If the parasitic element is a half-wavelength, it will act as a pure resistance. If the two elements, both a half-wavelength long, are very close together, the current in the parasitic element will be  $180^\circ$  out of phase with the current in the driven element, as is the case in the secondary and the primary of untuned transformers. This is normal transformer action. It has the same

effect as though both elements were excited  $180^\circ$  out of phase.

If the two elements are a quarter-wavelength apart, then the radiated energy reaching the parasitic element will be  $90^\circ$  behind that at the driven element, due to the distance traveled. While the induced current is  $180^\circ$  out of phase with the radiated energy inducing it, it is, in effect, leading the current in the driven element by  $90^\circ$ . It has the same effect as feeding the parasitic element with an excitation leading that of the driven element by  $90^\circ$ . The radiation pattern, then, will effectively be that of Figure 13-27B. Since the radiation is strongest in the direction of the driven element, the parasitic element is called a reflector.

If the two elements are placed closer together, the voltage induced in the parasitic element will be greater. For a reflector, maximum gain over a half-wave antenna alone is obtained with a spacing of 0.15 wavelength; for a director; 0.1 wavelength.

When a driven element and a parasitic element are placed 0.15 wavelength apart, as shown in Figure 13-30A, the induced voltage in R (parasitic element) lags that in the driven element by 0.15 wavelength, or  $54^\circ$ . This induced voltage is that acting at the center of R causing the parasite to be excited as though by a transmission line. If the radiation field due to the induced current is to reach the driven element in phase with the field leaving that element, there must be an additional phase shift. If the reflector (R) is slightly longer than a half-wavelength, it will present an inductance to the transmission line (in this case the induced voltage), and the current will lag the voltage. The amount of lag is varied by adjusting the length of the element. If the lag is properly adjusted, the reradiated voltage from R (Figure 13-30A) reaches A in phase and the radiation pattern will be as shown in the polar diagram.

The field passing from the reflector cuts the driven element and induces a voltage in it. This voltage changes the input current. The input impedance, which is a function of this current, is less than the 73 ohms for the antenna alone. The actual input impedance depends upon the diameter of the elements and the spacing, shape, and length of the reflector. It is usually about 50 ohms.

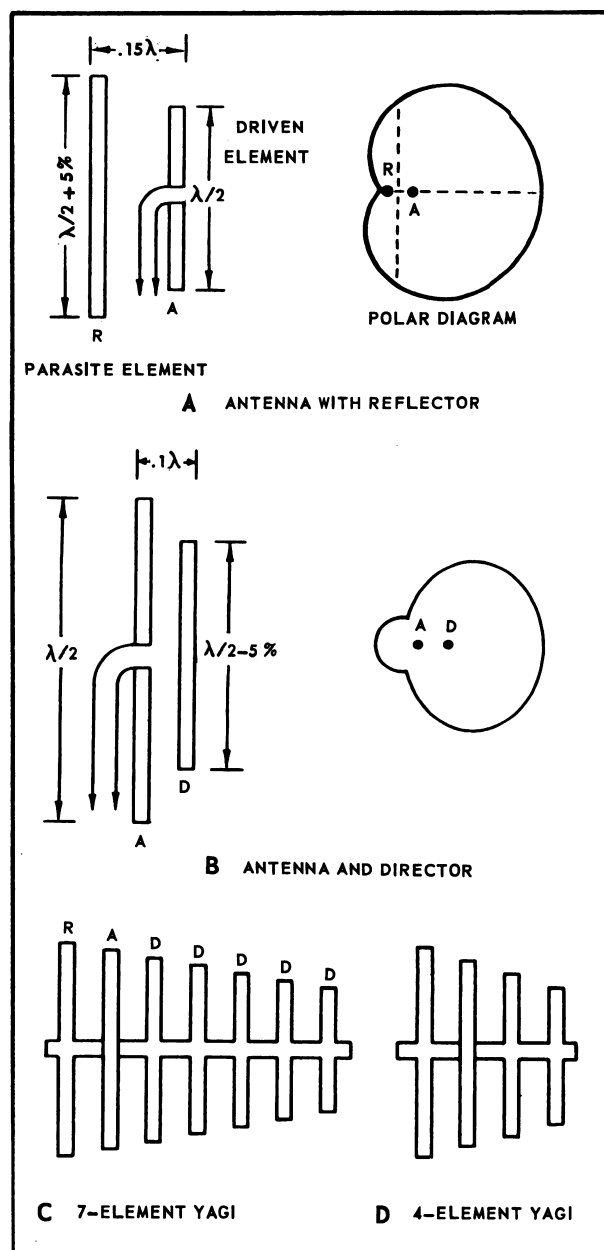


Figure 13-30. Parasitic Array

If the parasitic element is made shorter than the driven element and placed parallel to it with a spacing of 0.1 wavelength or slightly more, it will act as a director (Figure 13-30B). Most of the radiation will be in the direction of the parasitic element.

When this shorter parasitic element is placed 0.1 wavelength from the driven element, the leading effect caused by the capacitive action of the short element makes the reradiated energy from

the parasitic element reach the driven element approximately  $180^\circ$  out of phase with the field now leaving the antenna. This will produce cancellation in the direction of the driven element and maximum radiation in the direction of the director. Like the reflector, a director decreases the radiation resistance of the antenna.

If the spacing between the driven element and the director is increased, the director must be cut shorter to provide the proper phase shift; the farther away, the shorter the element.

### Directivity

The conventional method for describing the directivity of an antenna array is in terms of either the ratio of the power in the best lobe to the power radiated by a simple half-wave antenna, or the ratio of the power in the direction of the best lobe to the power in the opposite direction. The direction of the best lobe is usually called *forward* direction, while that in the opposite direction is called the *backward* direction. In other words, this second ratio is the power in the forward field to the power in the backward field. For example, the front-to-back ratio of the antenna array illustrated in Figure 13-30B is about 5:1. In decibels, this represents approximately a 7-db gain. The ratio between forward power and power from a single half-wave antenna is about 4 db.

The length and spacing of antenna elements which give maximum forward gain do not provide maximum attenuation in the opposite direction. In order to have the highest front-to-back ratio the tuning or length of the parasitic element must be adjusted. A reflector, for example, is tuned slightly lower or made slightly longer than for maximum gain. Adjusting the director for maximum attenuation of the rear lobe can increase the front-to-back ratio from about 5.5 to 17, with a decrease in the gain (compared to a simple half-wave antenna) of about 1 db.

Comparisons between the more directive types of arrays are made in terms of the beam angle. This is the angle between half-power points in the main lobe. In arrays the half-power points are the points where the electric field strength is 0.707 times as great as that along the axis of the beam. The narrower the beam, the more directive it is considered.

### Yagi Array

A comparison of two-element parasitic arrays will show little difference in gain between that using a reflector and the one using a director. However, by combining a driven element with both reflector and director, a more directional array can be built. Such an array is called a *Yagi* array. Yagi arrays are also designed with more than one director. Additional reflectors are not used since there is little radiation field behind the single reflector. Spacings are the same as for single elements — the reflector is usually 0.15 wavelength behind the driven element, and the director 0.1 wavelength ahead of the driven element for maximum gain in the forward direction.

Figures 13-30C and D show two Yagi arrays, one of seven elements, the other of four. The driven element (A) is insulated from the support, but the reflector and directors are all welded to this piece of tubing running parallel to the direction of propagation. Chapter 15 contains more on Yagi arrays.

The gain of parasitic arrays is difficult to calculate or to measure. However, based on ground measurements and actual experience, the average gain for a 3-element Yagi is about 6 to 7 db.

One of the disadvantages of the Yagi array is that tuning of the elements is critical. If the antenna is used for frequencies other than the one for which it is designed, gain, front-to-back ratio, and radiation pattern change completely.

When used with the frequency for which it was designed, the antenna shown in Figure 13-30D has a beam width of about  $50^\circ$ . The 7-element Yagi in Figure 13-30C has a beam width of about  $19^\circ$ .

### Folded Dipole

In multielement arrays the input impedance drops as low as 15 ohms. In these arrays special matching devices are needed for matching this low impedance to the higher impedance of most r-f lines. One method for making the match is to use the Delta matching system previously described. Another method, the one most often used in radar equipment, uses the folded dipole as shown in Figure 13-31.

A folded dipole is a full wavelength conductor that is folded to form a half-wave element. A better description is that it consists of a pair of

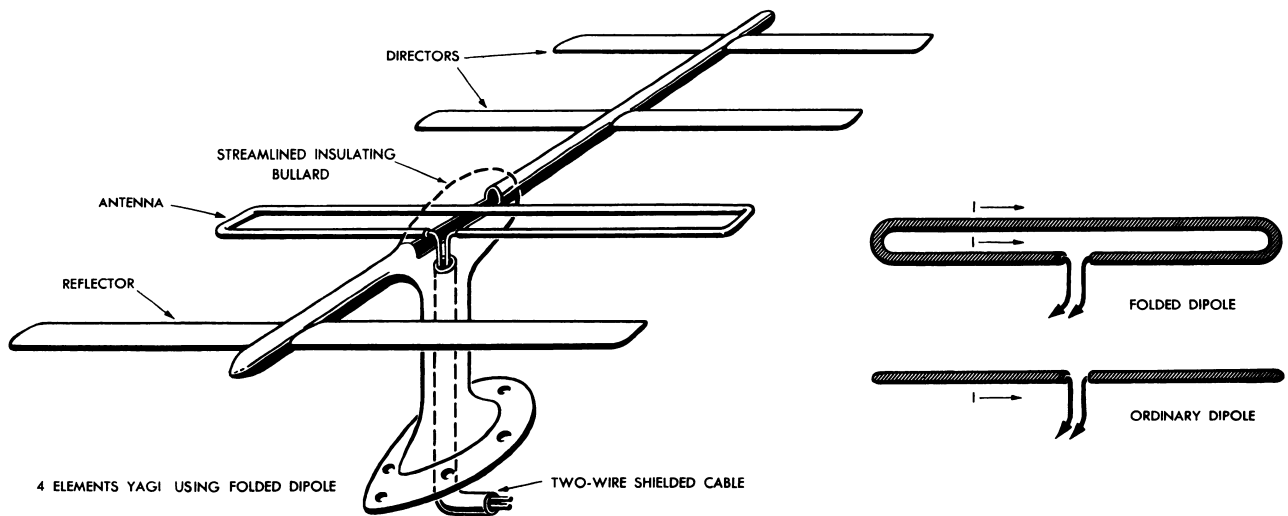


Figure 13-31. Folded Dipole

half-wave elements connected together at the ends. In it the voltage at the ends of each element must be the same. In operation the field from the driven element induces a current in the second element. This current is the *same* as the current in the driven element.

An ordinary dipole with a given current  $I$  produces a certain field intensity in space. Due to this field, there is also a certain power density per square meter in space. This power density is produced by the input power  $P$ . The relationship between the input resistance, the current, and the input power is expressed by the equation,

$$R = \frac{P}{I^2}$$

Another consideration with the folded dipole is that when the same current  $I$  exists in each of the two sections, the field strength in space is doubled. This causes the power density per square meter to increase four times. In turn, the input power must be four times as great. In this case to balance the equation, it is necessary to multiply  $R$  by 4 as follows:

$$\frac{4P}{I^2} = 4R$$

As long as each section of a dipole has the same diameter, the input resistance is four times that of the simple (half-wave) dipole. Increasing the diameter of one section makes the increase in impedance still greater. While the input impedance to the driven element of a parasitic array

drops to about a fourth of the value of the coaxial cable impedance, the use of a folded dipole increases the impedance by about four times. In this way a good impedance match is effected.

## REFLECTORS

### Parabolic Reflectors

A spherical waveform spreads out as it travels because the energy travels in a direction perpendicular to the wave front. This produces a pattern that is not too sharp or directive. On the other hand, a plane wave front does not spread out because all the wave front moves forward in the same direction. For a sharply defined radar "beam" we need to change the spherical wave front from the antenna into a plane wave. A parabolic reflector is one way of doing this.

As the transmitting frequencies become higher, the wavelength shorter, radio waves become more like light waves. Microwaves (with wavelengths measured in centimeters) travel in straight lines as do light rays. They can be focused and reflected just as light rays can.

A metallic parabolic reflector is used to focus light rays in an automobile headlight. The light source is placed at the focal point. Light rays striking the parabolic reflector are reflected at angles that result in parallel paths for all the light rays. The same thing can be done with microwaves. While light rays require a highly

polished surface, microwaves can be reflected by a large metal surface.

In Figure 13-32A a point radiation source is placed at the focal point F. The field leaves this antenna with a spherical wave front. As each part of the wave front reaches the reflecting surface, it is shifted 180° in phase and sent outward at angles that cause all parts of the field to travel in parallel paths. Because of the shape of a parabolic surface (paraboloid), all paths from F to the reflector and back to line XY are the same length. Therefore, all parts of the field arrive at line XY the same time after reflection.

If a dipole is used as the source of radiation, there will be radiation from the antenna out into space as well as toward the reflector. This energy that is not directed toward the paraboloid has a wide beam width which would destroy the narrow

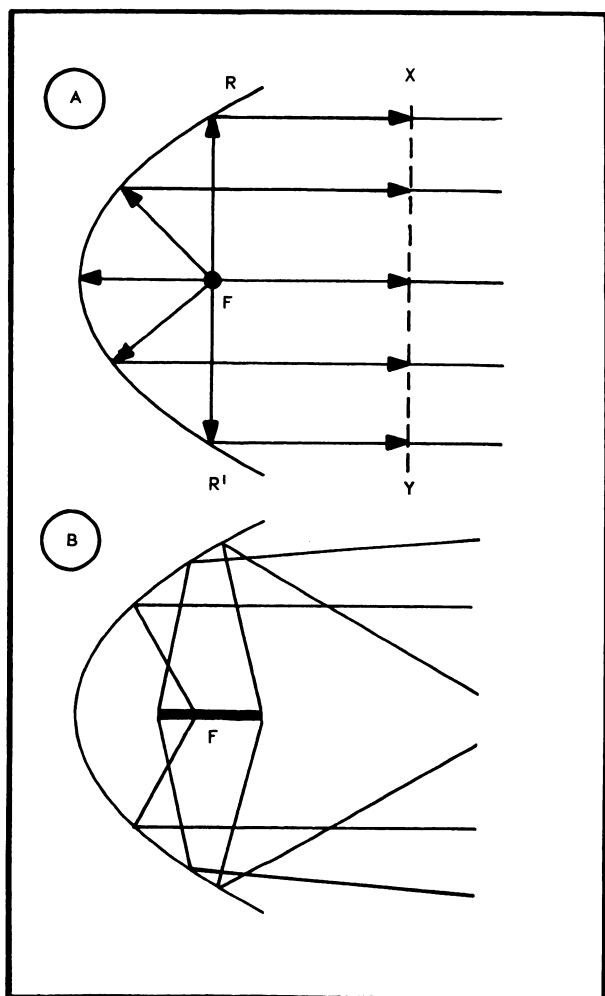


Figure 13-32. Parabolic Reflector

pattern from the parabolic reflector. To prevent this, a parasitic reflector may be used with the half-wave driven element. In this way all of the radiated energy is directed into the parabolic reflector, and a narrow beam is produced.

If part of the antenna is outside the focal point, energy from that part will not be reflected parallel to the rest (Figure 13-32B). For this reason, the antenna is often physically shorter than a half-wave. For the radiating element to approximate a point source, the parabolic reflector should be at least ten times the size of the dipole being used.

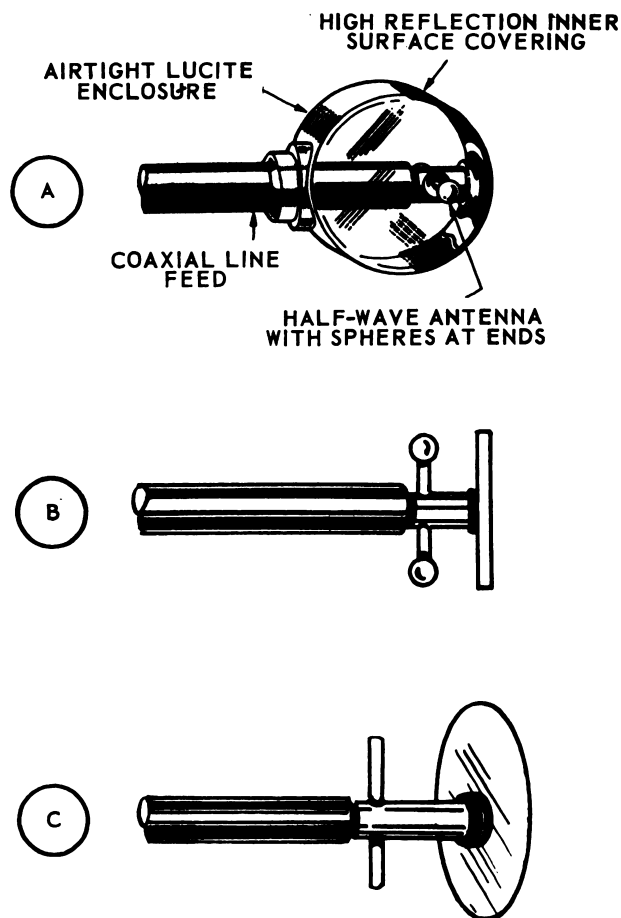


Figure 13-33. Dipole for Exciting Parabolic Reflector

Figure 13-33A shows the exciting antenna for a parabolic reflector for a 10-cm radar system. The dipole is mounted a quarter-wave back from the short on the coaxial line. To sharpen the focal point the antenna is physically less than a half-wave long. However, the balls at the end

make it electrically a half-wave long. This broadens the band of frequencies it will handle. The airtight cylinder in which the antenna is enclosed permits the coaxial line to be pressurized. The inner surface of half of the cylinder, that is, the side away from the parabolic reflector, is coated with a reflecting foil. This reflecting surface directs energy from that side of the antenna into the parabolic reflector. Without this reflector half of the antenna radiation would be nondirectional.

The natural directivity of a dipole causes the pattern from a parabolic reflector to be somewhat sharper in the plane containing the dipole than in the other plane. For this reason the dipole is mounted horizontally for maximum azimuth accuracy in radar systems. If vertical accuracy is of primary importance, the dipole is mounted vertically.

Figure 13-33B shows the use of a parasitic reflector with the dipole; Figure 13-33C shows a disc that is placed in front of the dipole.

There are many forms the parabolic reflector may take. The larger the reflector with respect to the wavelength, the narrower the beam width.

This fact makes it possible to shape a beam by changing the dimensions of the reflector.

### Truncated Paraboloid

Figure 13-34A shows a truncated paraboloid. Since the reflector is parabolic in the horizontal plane, the energy is focused into a narrow beam. Since the reflector is truncated, or cut, so that it is shortened vertically, the beam spreads out vertically instead of being focused. Such a fan-shaped beam is used to determine azimuth accurately. Since the beam is wide vertically, it will detect aircraft at different altitudes without changing the tilt angle of the antenna.

The truncated paraboloid reflector can also be used in height-finding equipment. Since the reflector is parabolic in the vertical plane, the energy is focused into a narrow beam; since the reflector is truncated, or cut, so that it is shortened horizontally, the beam spreads out horizontally instead of being focused. Such a fan-shaped beam is used to determine elevation very accurately. Since the beam is wide, a greater area can be scanned in a shorter time than is possible with a narrow, pencil-like beam.

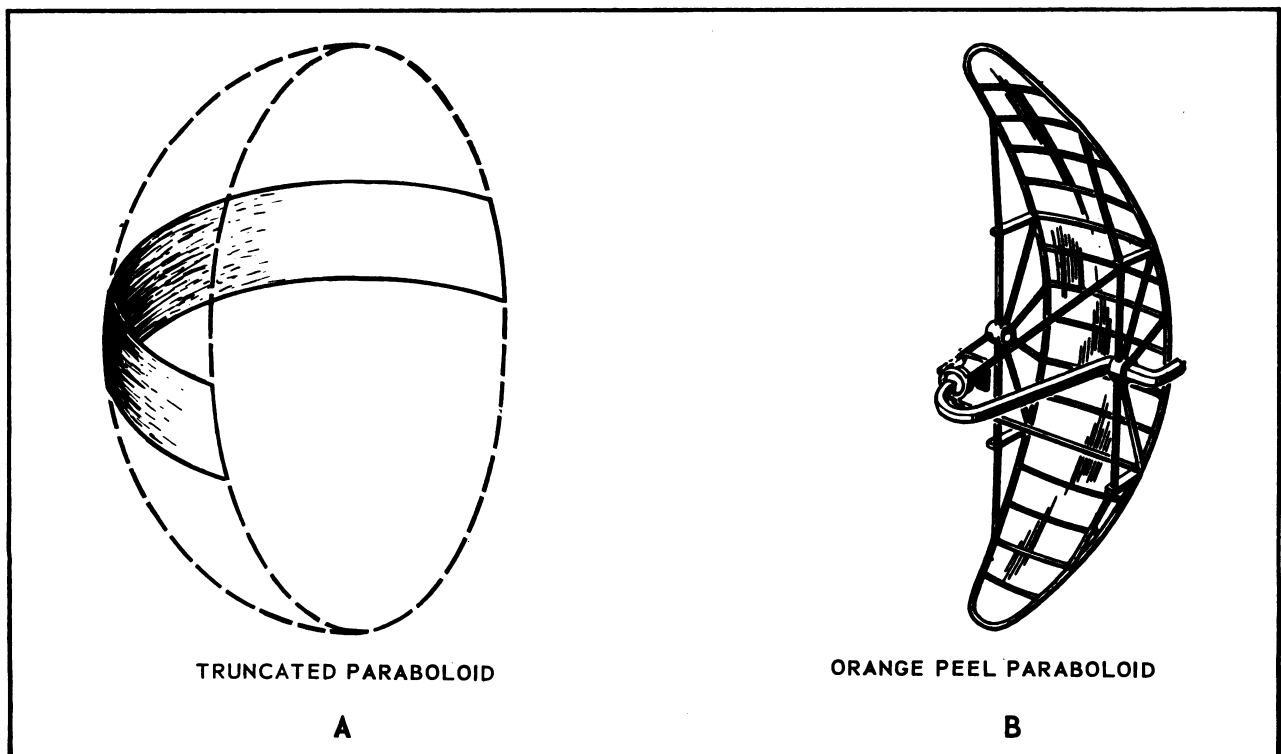


Figure 13-34. Paraboloid Reflectors



### Orange-Peel Reflector

A section of a complete circular paraboloid, often called an *orange-peel* reflector because of its shape, is shown in Figure 13-34B. Since the reflector is narrow in the horizontal plane and wide in the vertical, it produces a beam that is wide in the horizontal plane and narrow in the vertical. In shape it resembles a huge beaver tail. The r-f energy is sent into the parabolic reflector by a horn radiator fed by a waveguide. The horn nearly covers the shape of the reflector, so almost all of the r-f energy is directed into the reflector, very little escaping at the sides. This type of antenna system is generally used in height-finding equipment.

### Cylindrical Paraboloid

When a beam of radiated energy that is noticeably wider in one cross-sectional dimension than in the other is desired, a cylindrical paraboloidal section approximating a rectangle can

be used. Figures 13-35A and 13-35B illustrate this antenna. A parabolic cylinder has a parabolic cross section in one dimension only; therefore, the reflector is directive in one plane only. The cylindrical paraboloid reflector can be fed by a linear array of dipoles, by a slit in the side of a waveguide, or by a thin waveguide radiator. Rather than a focal point, this type of reflector has a series of focal points forming a straight line. Placing the radiator, or radiators, along this focal line produces a directed beam of energy. As the width of the parabolic section is changed, different beam shapes are obtained. This type of antenna system is used in search and in Ground Control Approach (GCA) equipments.

### Cosecant-Squared Reflector

Some antennas are designed to produce special pattern shapes or to produce a nonuniform distribution of radiated energy. Airborne equipment has a special need for such equipment. In

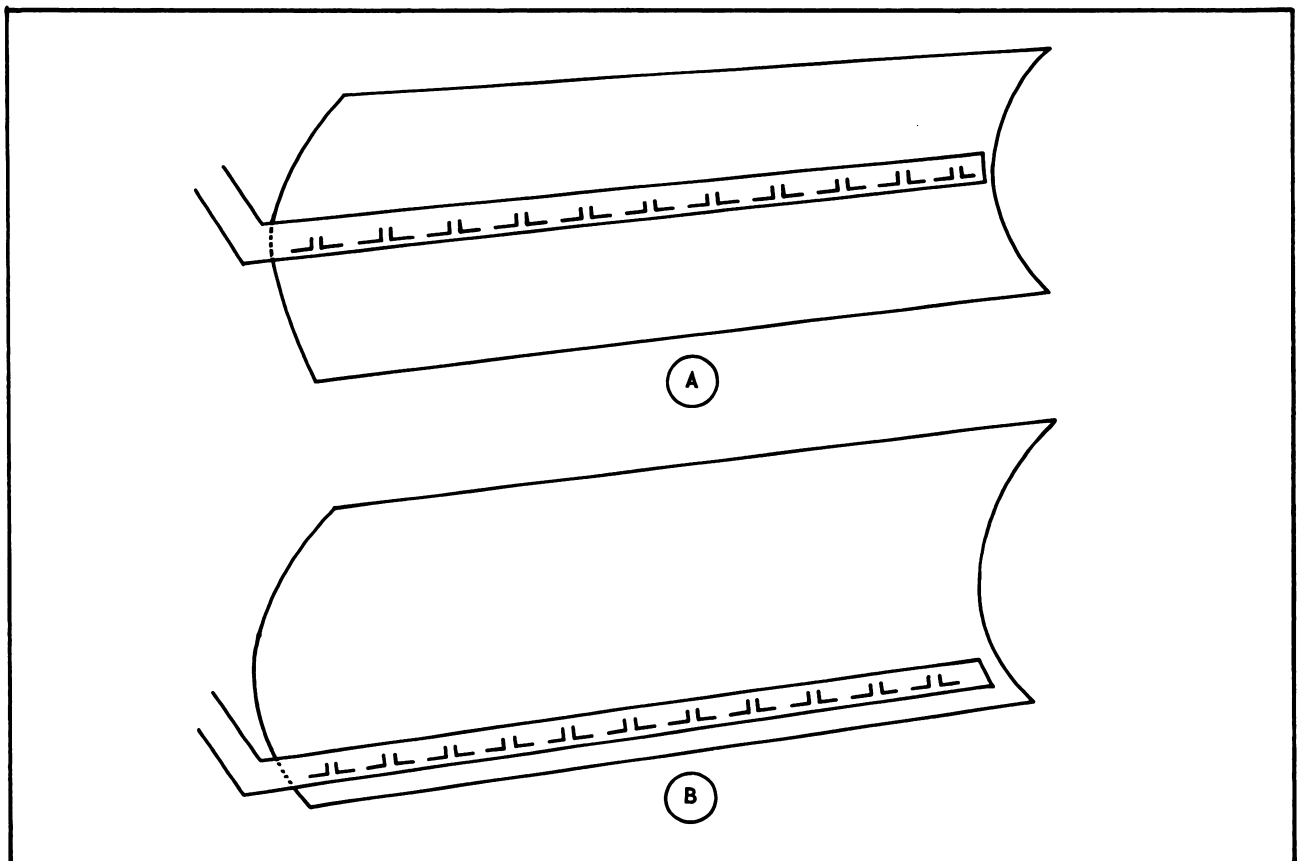


Figure 13-35. Cylindrical Paraboloid

scanning the surface of the earth in a forward direction, the surface some distance ahead is the area of importance. Normally this area would receive and return the weaker signal. The antenna, then, must be designed to reduce the energy that would be radiated downward and concentrate it in a forward direction.

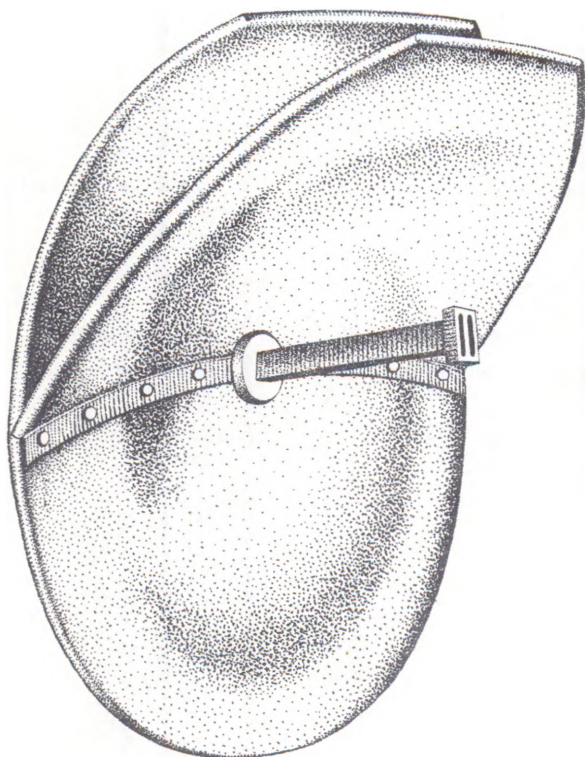


figure 13-36. Cosecant-Squared Reflector

If the top of the parabolic reflector is bent forward, the desired pattern will be produced. Rather than change the actual shape of the parabolic reflector, a shaped piece of metal, known as the cosecant-squared *skirt*, is placed inside the upper part of the reflector. The signal strength now varies directly as the cosecant of the angle between the horizontal and the direction of the antenna. Since the power varies directly as the square of the field intensity, this antenna reflector is called a cosecant-squared reflector. The cosecant-squared principle can be used on equipment other than airborne; however, Figure 13-36 shows a type of cosecant-squared reflector used on a piece of airborne equipment.

#### Conical Scanning with Parabolic Reflector

Conical scanning is used in radar equipment to direct antiaircraft fire and in automatic tracking systems. It is possible to obtain angular precision of only a few hundredths of a degree. In this type of scanning a point on the radiated beam forms a circle at the base of the cone, as illustrated in Figure 13-37. The axis of the beam forms the cone. When a target is within the cone but off the axis, it will receive a different amount of energy as the beam makes its conical scan. As the target approaches the axis of the cone, variations of reflected energy become less and less. The reflected power does not vary when the target is on the axis of the cone. By detecting the amplitude variations of the received pulses, an

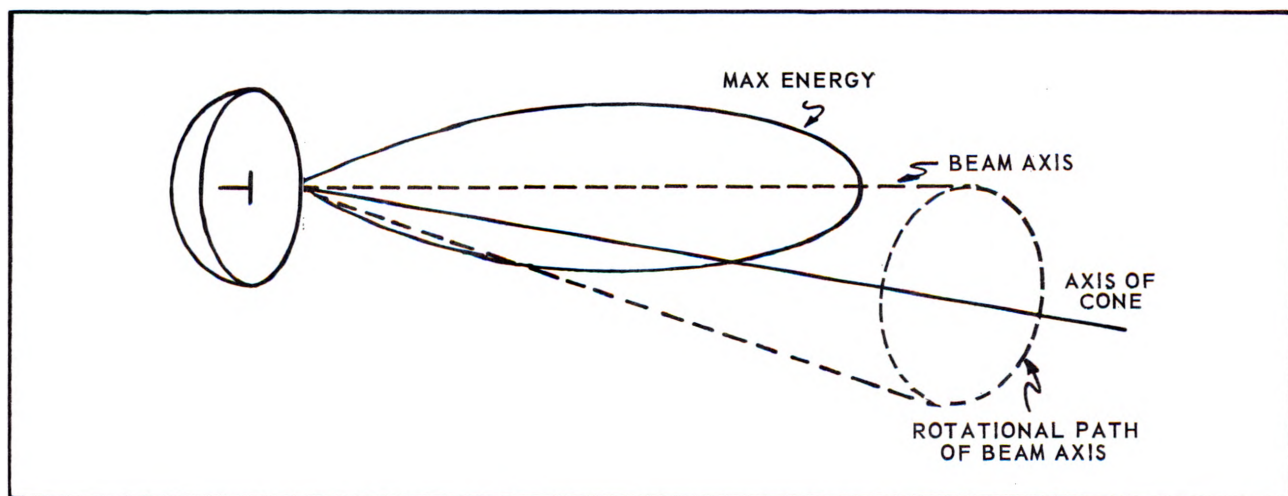


Figure 13-37. Conical Scanning

automatic tracking system is made possible. Conical scanning may be produced by rotating the dipole antenna around the axis of the parabolic reflector.

## TYPICAL AIRBORNE ANTENNA SYSTEMS

### General

The form and size of aircraft antennas are dependent upon their use. One of the prime considerations is that the antenna must not interfere with the flight of the aircraft. Normally the antenna is located away from all dielectric obstructions, so that nearby objects will not absorb any radiated energy. The ground for the antenna system is practically perfect as it is the metallic skin of the aircraft. Short transmission lines with their inherent lower loss can be used as the antenna is usually near its associated equipment.

The prop-driven aircraft that are still being used have fixed wire or stub-mast antennas for communications. Some are flush-mounted into the skin of the aircraft and some are mounted in the tail section. They are located so that the antennas offer minimum drag to the aircraft during flight.

Because of the speed of jet aircraft, the communications antennas used are different from

those used in prop-driven aircraft. UHF, VHF, and HF antennas are a part of the rudder section, located at the top of the rudder. The physical size of the antennas is determined by the wavelength of the frequencies used. Each antenna is insulated from the others. Installed in this manner the antenna offers no additional drag to the aircraft. Where other communications antennas are needed, they are flush-mounted, generally in the nose of the aircraft.

### Navigational Antenna

The AN/APN-69 is an Airborne Radar Beacon set useful in navigation. This set has found wide use in mid-air refueling operations. It was designed and constructed to be used with 3-cm Airborne Search Radar equipment. For successful operation of the AN/APN-69 there must be a search radar set operating in the vicinity as a beacon-interrogator.

The antenna, as shown in Figure 13-38, serves to receive radio-frequency energy from the interrogating aircraft or radiate r-f energy to the interrogating aircraft. The antenna is designed to provide omnidirectional coverage. The radiating element, which protrudes outside of the aircraft structure, is covered by a radome made of a nylon cloth shell filled with a foam material. The antenna is installed in a fixed position and does not require operating adjustments.

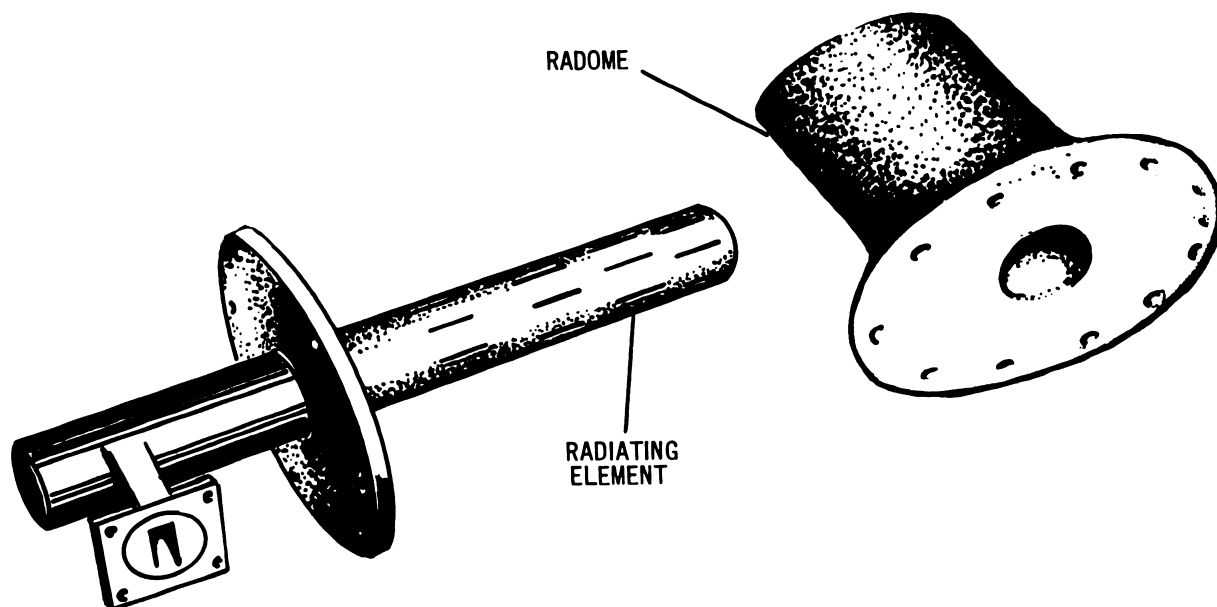


Figure 13-38. Antenna and Radome AN/APN-69



## Airborne Search Radar

Radar Set AN/APN-59 is a small, lightweight radar system which operates as a navigational and search radar, a weather observation radar, or radar beacon interrogator-receiver.

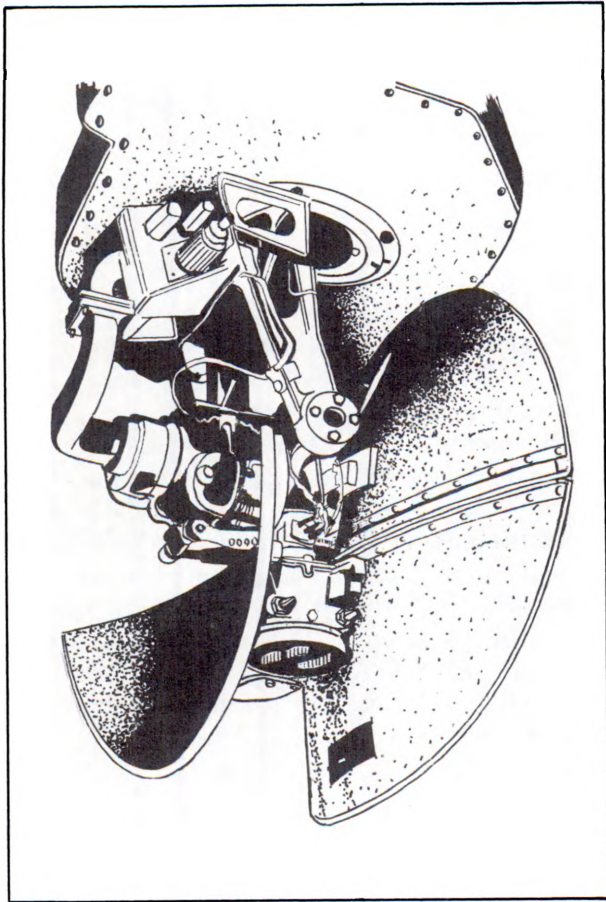


Figure 13-39. Antenna Reflector AN/APN-59,  
Right Side Detail

When used as a search radar, the system displays a visual, map-like scope picture. When used as a weather radar, it displays less substantial objects such as storm fronts and heavy rainfall. When used for radar beacon navigation, it transmits an interrogating signal and then displays, in plan position, the space-coded identification of the automatic radar beacon reply or replies. As the aircraft pitches and rolls, any displayed signal is prevented from shifting position or disappearing by the use of a gyro-stabilized r-f beam pattern.

The transmitted signal may be radiated as either a fan or a pencil beam. The fan beam is shaped to sweep out at a vertical angle of nearly  $60^\circ$  downward from the level of flight with a power pattern that returns echoes of nearly equal strength from remote or nearby targets. The pencil beam is similar to a searchlight beam. Tilt is provided so that the beam may be directed downward or slightly upward, as required.

The antenna may be set to rotate about an axis perpendicular to the deck plane of the aircraft, or gyro-stabilized to rotate about an axis perpendicular to the earth. The antenna may be driven in full rotation either clockwise or counter-clockwise, or stopped; and it may be set to scan a sector across the aircraft heading. The antenna has two speeds of scan that are chosen automatically by the function and range in use. Both an 18-inch and a 30-inch reflector and feed assembly are available. The 30-inch reflector gives better definition due to the narrow horizontal beam width of  $3^\circ$ . This larger reflector will be used where space and weight are not at a premium.

The waveguide in the antenna is all of the rigid type. This requires three rotating joints: (1) azimuth, (2) level (pitch), and (3) cross level (roll). Between the level and cross-level joints of the 18-inch reflector on the top side of the waveguide is a machined surface for placing level test instruments. Figure 13-39 shows the 18-inch and 30-inch reflector units of the AN/APN-59.

The antenna shown in Figure 13-40 is used with the AN/APS-42A, an airborne search set. The illustration shows a portion of the r-f waveguide ① which conducts the transmitter pulses to the antenna and the received pulses from the antenna, the 18-inch paraboloid reflector ② and spoiler grid ③ which shape the transmitted beam into the desired radiation pattern, the azimuth drive motor ④ that moves the antenna on full or sector scan at one of two speeds. A gyroscope ④ functions with associated circuits to stabilize the antenna when the aircraft rolls or pitches. The antenna tilt system ⑤ enables the operator to tilt the antenna from  $10^\circ$  above the horizontal plane to  $15^\circ$  below the horizontal plane for additional coverage. The pattern change system makes possible the use of two radiated patterns: pencil or obstacle beam, and mapping beam.



## AEW Equipment

The largest antennas used in airborne equipment are those used in the Aircraft Early Warning Radar System. The aircraft used for this system is the RC-121D, shown in Figure 13-41. Two unclassified pieces of equipment used in an airborne early warning system are the AN/APS-45, height finder, and the AN/APS-95, search set. Figure 13-41 shows the location of the AN/APS-45 antenna on the aircraft.

The height-finder antenna for the AN/APS-45 is shown in Figure 13-42. The widest part of the orange-peel reflector is about 4 feet; the height is about 9½ feet.

This antenna provides for focusing, radiating, and receiving the radar pulses, and for conveying target azimuth and elevation information to the control-indicator. The beam position is stabilized in elevation with respect to line-of-sight, and in azimuth, with respect to north. The antenna can be rotated continuously clockwise or set to any desired azimuth position. In addition it can be set to any desired elevation angle within the 16° limits of the elevation sector scan.

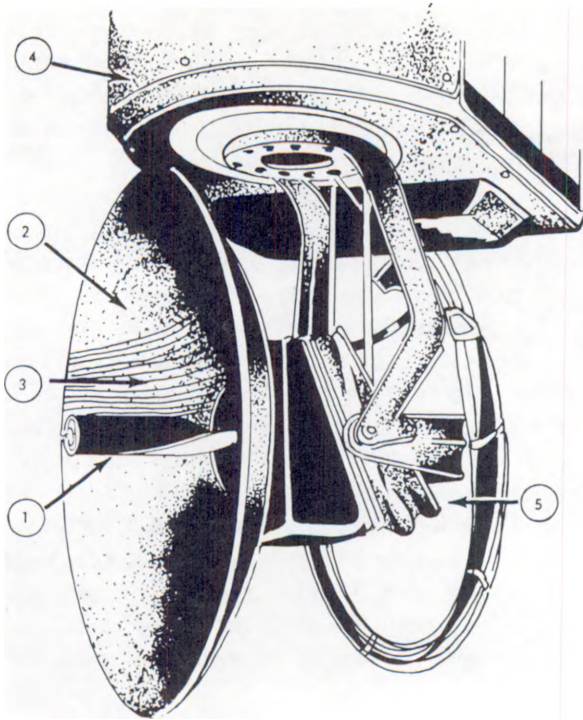


Figure 13-40. Antenna Unit AN/APS-42A

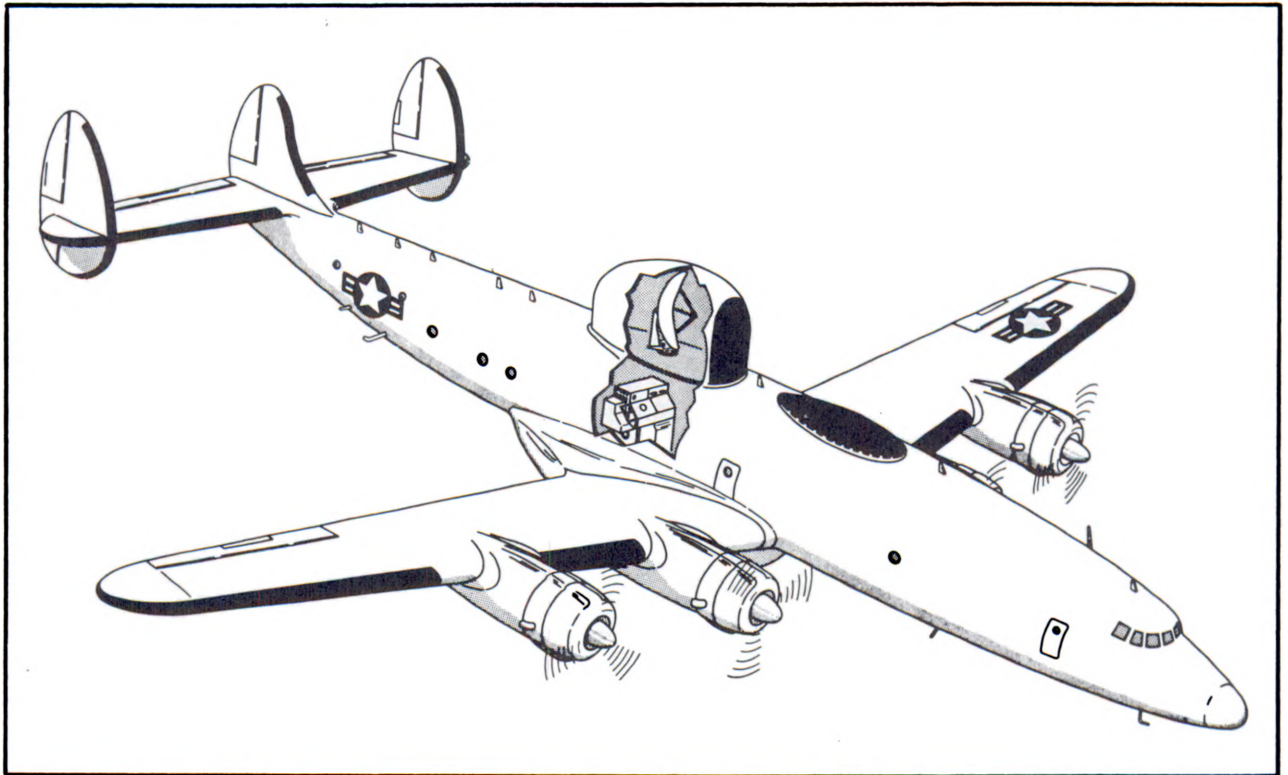


Figure 13-41. RC-121D AEW Aircraft



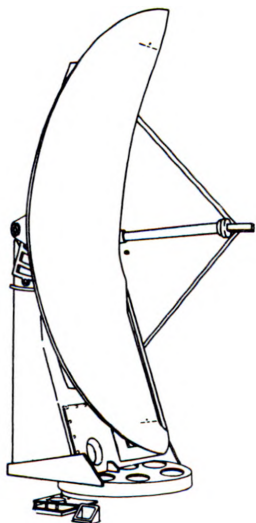


Figure 13-42. Antenna for AN/APS-45

The search set antenna, Figure 13-43, functions to radiate radar and IFF transmitted power, and receive radar and IFF return signals. The radar portion of the antenna consists of a 12-element linear array of bow-tie-slot resonant cavities. Each element consists of a rectangular cavity containing one or two bow-tie slots in the front panel. Each bow-tie is fed by a rigid coaxial feed; this, in turn, is fed by a rigid

coaxial corporate feed. The corporate feed is fed by a hybrid power divider which contains a sum and difference channel. The sum channel is used during both transmission and reception, while the difference channel is used during reception only.

The antenna is mounted to the rotating portion of the pedestal casting which protrudes through the aircraft fuselage, so that the antenna is mounted below the aircraft. The rotary joint provides continuity between the fixed and rotating sections of the antenna pedestal. Rotation in either direction is possible at a speed of 2, 4, or 6 rpm. The antenna assembly is protected by a weatherproofed radome housing. The physical length of the antenna is approximately 17 feet and the height is approximately 4 feet.

A special feature of the APS-95 is that it uses the principle of a Displaced-Phase-Center Antenna. The function is to initiate phase relationships in order to cancel off-axis clutter in the MTI unit.

The IFF antenna is a 2-dimensional linear array of two rows of 24 identical slot-fed dipoles. They are vertically positioned and each pair is covered with a radome for protection. This is illustrated in Figure 13-43. The IFF antenna uses the front panel of the radar antenna as a reflecting surface.

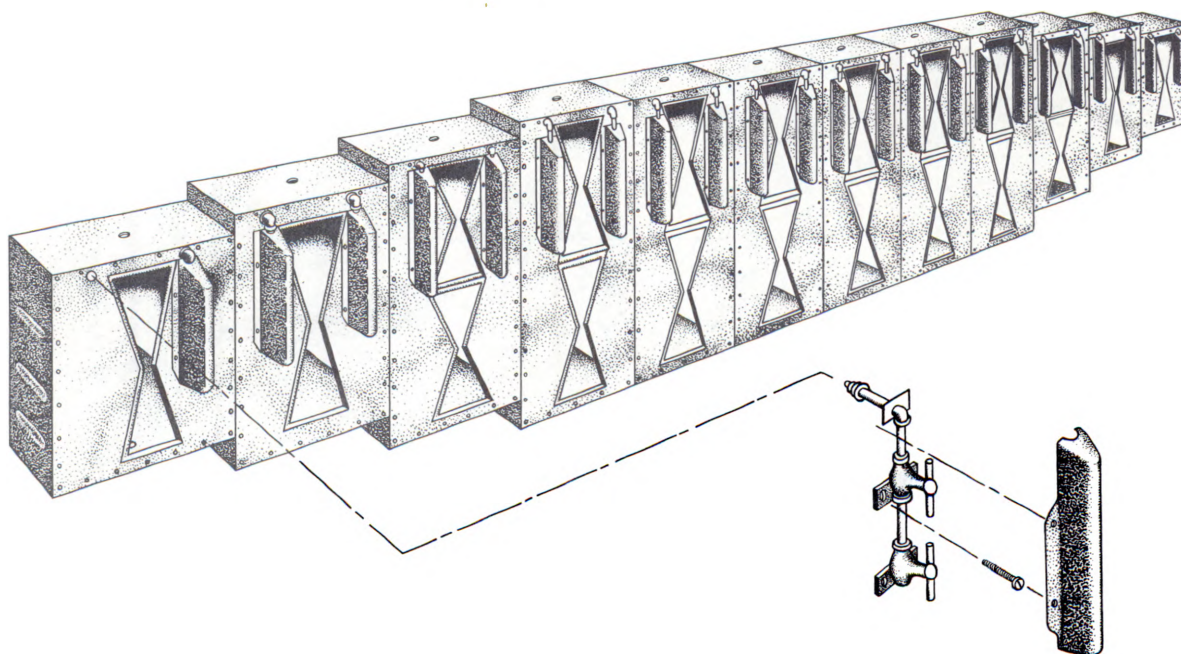


Figure 13-43. Antenna Assembly AN/APS-95



## GROUND RADAR ANTENNAS

### General

The effectiveness of any radar set is influenced to a great degree by the nature and quality of the radar's antenna. The range, accuracy, and resolution of the set is affected by the electrical properties of the antenna. The directional features of radar antennas makes it possible to illuminate certain selected targets and to receive the returns from these same targets for any given time. Directivity allows all of the energy to be concentrated in one general direction. Radar antennas are as numerous in kind as is radar equipment. They must be designed according to the character and particular function of the radar.

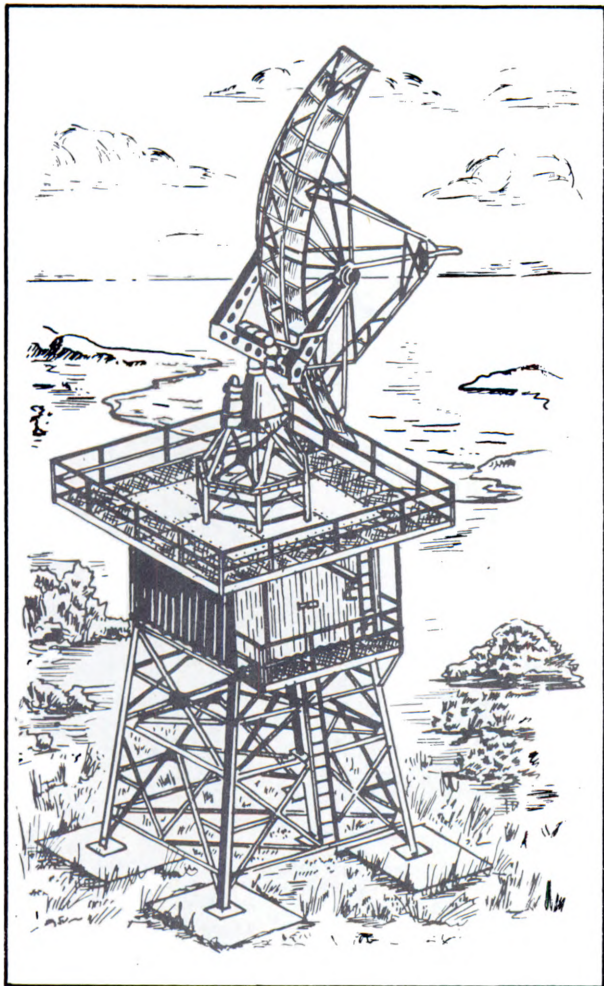


Figure 13-44. AN/FPS-6 Antenna

### Height Finder Radar

The AN/FPS-6 is an air-transportable, high-power, long-range, fixed-station height-finder radar used in aircraft control and warning. AC&W systems consist of ground search radar equipment combined with height-finding equipment. The AN/FPS-20 and the AN/FPS-6 form such a combination. The FPS-6 radar set in conjunction with the FPS-20 enables the aircraft controller to place the enemy aircraft accurately in azimuth, range, and height. This three-dimensional picture allows the controller to easily vector the intercepting aircraft to the point of interception.

The reflector of the FPS-6 antenna is a truncated orange-peel type. The reflector is horn fed, has a height of 30 feet and is 7½ feet in width. The reflector nodding scan is from  $-2^{\circ}$  to  $+32^{\circ}$  with a nodding frequency of 20 or 30 npm. The antenna has a  $360^{\circ}$  azimuth bearing capability; it can operate while rotating continuously or at a stationary position at any azimuth. Figure 13-44 shows an AN/FPS-6 antenna on a fixed tower.

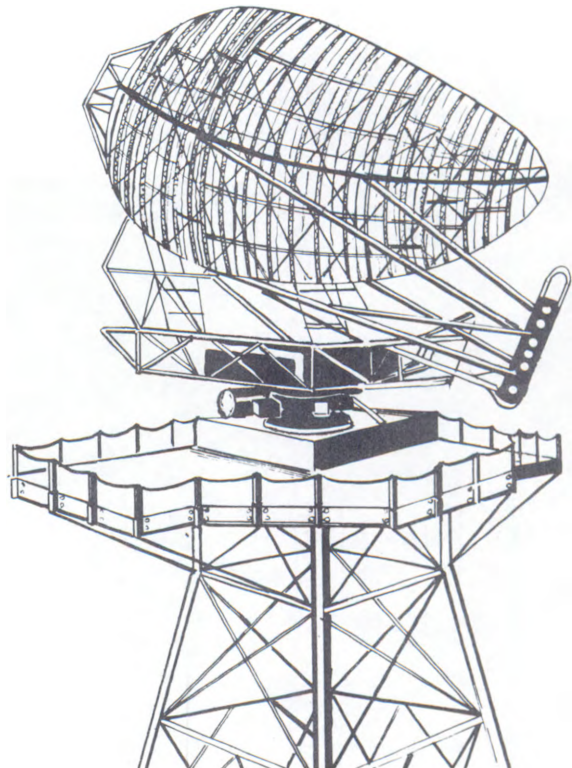


Figure 13-45. AN/FPS-20 Antenna

## Ground Search Radar

The AN/FPS-20 is a long-range, high-powered ground search set. It furnishes a very important addition to the Air Defense system by providing a range-azimuth coverage. In conjunction with height and identification information, this completes the AC&W system.

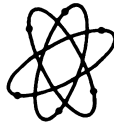
In a rotating antenna system, such as used in the AN/FPS-20 shown in Figure 13-45, a transition must be made between the fixed waveguide system below the antenna pedestal and the rotating waveguide system above it. The FPS-20 uses a rotary joint in the antenna pedestal as the transition from fixed to rotating waveguide.

A high-powered, three-cavity klystron is used as a power amplifier in the transmission system. The desired antenna beam pattern coverage of only one beam is accomplished with this type of transmission and a line feed reflector. The energy pattern is narrow in the horizontal plane

and very broad in the vertical plane. Such a beam gives good azimuth definition and also detects high flying targets.

The azimuth rotation system rotates the antenna through  $360^\circ$  of azimuth search. The pedestal, drive motor, and the azimuth data generator unit which is mounted on the pedestal is included in this system. The antenna reflector and the rotary coupler are mounted on the antenna bull gear, which is driven by a 20-hp three-phase motor through a gear reducer. The antenna rotates at speeds of 3.3, 5.0, 6.6, and 10.0 rpm.

The reflector shape is a section of a paraboloid rotated around its axis. The outline of the reflector is an ellipse; the long dimension of the reflector is approximately 40 feet long, and the short dimension is approximately 16 feet long. Figure 13-45 shows the AN/FPS-20 antenna mounted on a fixed tower.



# Synchros and Servomechanisms

Up to now we have analyzed a great variety of circuits used in electronics in general, and in a number of radar sets in particular. This chapter is to acquaint you with devices which transmit position between these circuits and remotely located gages, dials, and controls. The chapter discusses the basic theory of two of these devices — the synchro, a machine which converts mechanical position into electrical position, or electrical position into mechanical position; and the servomechanism, a system that amplifies and transmits mechanical position from one place to another by electrical means.

## USES

There are many operations in radar where it is desirable for two shafts to rotate in synchronization. But because the distance between the two shafts is too great, or because one shaft cannot develop enough torque to rotate the other one, it is not always feasible for the shafts to be connected together mechanically. Therefore, shafts which are designed to rotate in synchronization are generally connected electrically.

Electronic systems for rotating two shafts in synchronization are called *remote indication* or *remote control* systems. A system in which the torque of the shaft which controls the operation (the *control* shaft) is amplified before it is applied to the shaft that is connected to the load (the *load* shaft), is called a *servomechanism* or *servo* system.

As you can see by the name, a remote indication system is designed either to produce at one

position an indication of an operation at another position, or to transfer information from one position to another. Sometimes this system is called a *data transmission system*. Many radar antennas, for example, are designed to rotate 360° but are located in places where it is not possible for the operator to see them. Therefore, a data transmission system is employed to give the operator the direction in which the antenna is pointed.

A pointer attached to a shaft rotates in synchronization with a shaft connected to the antenna, and registers the position of the antenna on a dial at the operator's position. The devices through which the position of the antenna is transmitted to the dial are called *synchros*. They resemble small electric motors. The one at the antenna is called a *synchro generator* or *transmitter*; the one at the operator's position is called the *synchro motor* or *receiver*.

The word synchro, as you can see, is a contraction of the word *synchronous*. Trade names which are frequently used for these devices are *Selsyn*, *Autosyn*, *Magnesyn*, and *Synchrotie*.

In the remote indication system described, it is possible for the operator to rotate the antenna itself from his position. However, this requires more torque than can be transmitted efficiently through the system. To provide the necessary torque a servo system must be used. In the servo system, the operator turns a hand wheel (or some other control). This moves the shaft connected to the hand wheel out of step with the shaft connected to the antenna.

The out-of-step condition produces a voltage called the error voltage. This voltage is amplified by the torque amplifier and applied to a motor that rotates the antenna. The motor rotates in a direction that counterbalances the error produced by the out of step relationship between the hand shaft and the antenna shaft, and thus causes the antenna to take the same position as established by the hand control.

### POSITIONING A SHAFT WITH DC VOLTAGES

Although the synchro is an AC device, DC voltages are sometimes used for positioning shafts. Since an understanding of shaft positioning with DC will help you to analyze the operation of the synchro, it is well to take it up first.

There are a number of ways for positioning a shaft with DC voltages. One is to mount a bar magnet on the shaft and to locate it near a solenoid (coil). When a voltage is applied to the coil, the shaft will rotate to either of the two positions shown in Figure 14-1. When you reverse the polarity of the DC voltage, the magnet will likewise reverse its position. The position of the magnet shown is arbitrarily designated the  $0^\circ$  position. Furthermore, the positive angles represent rotation in the counterclockwise direction. In examining this arrangement closely you can see that only two positions of rotation are possible. Therefore, it would not be very satisfactory as a remote indicating system.

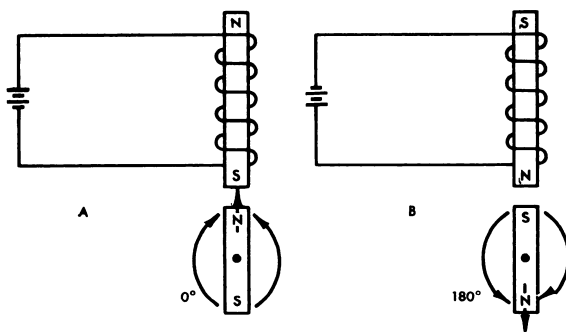


Figure 14-1. Using One Coil to Position a Magnet

Another system capable of rotating a shaft to more positions than a system using one coil, uses two coils that are placed at right angles about the magnet. When a single coil is placed as

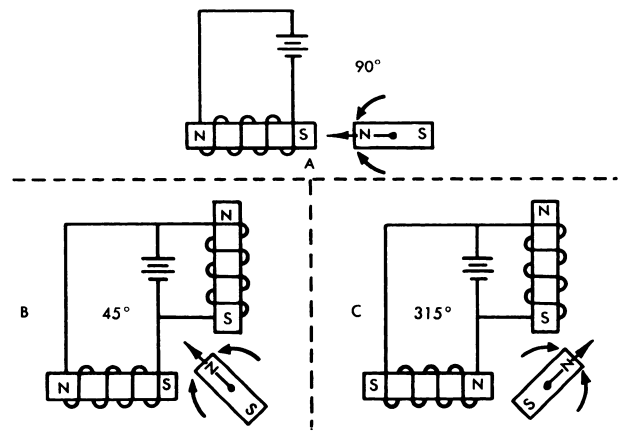


Figure 14-2. Using Two Coils Placed at Right Angles

shown in Figure 14-2A, the magnet will rotate to the  $90^\circ$  position; when the polarity of the coil is reversed, the magnet will rotate to the opposite, or  $270^\circ$  position. If, however, you connect two identical coils at right angles into the circuit (Figure 14-2B), the magnet will rotate to the  $45^\circ$  position. If you reverse the polarity of both coils, the magnet will reverse its position and come to rest at the  $225^\circ$  position.

Reversing the polarity of the coil in the horizontal position causes the magnet to assume the  $315^\circ$  position shown in Figure 14-2C. If the polarity of both coils in Figure 14-2C is reversed, the magnet will take a  $135^\circ$  position. Thus, you can see that by applying the same voltage to one or both of two coils that are placed

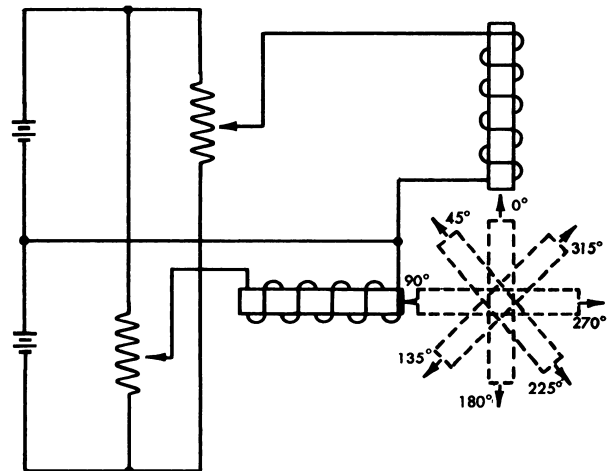


Figure 14-3. Using Two Coils and Potentiometers

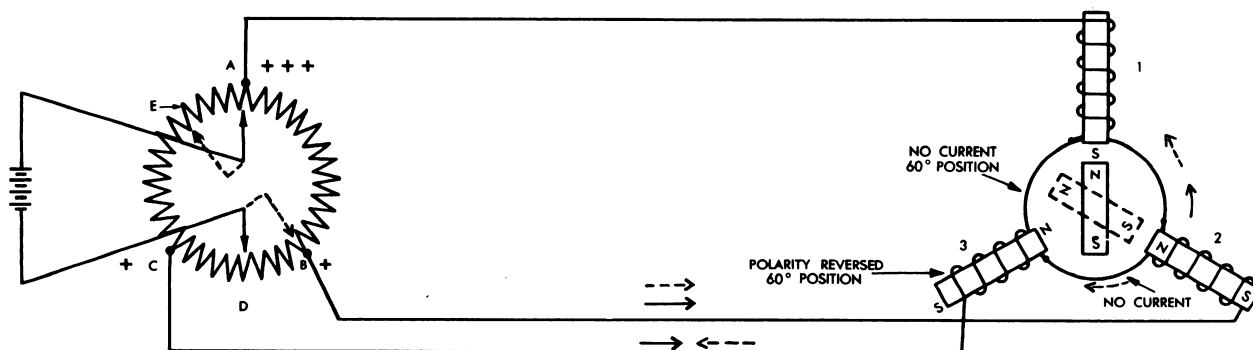


Figure 14-4. Potentiometer and Magnet Remote Indicator

at right angles, and by reversing the polarity of the coils, the magnet will take any one of eight positions.

A further method of positioning a shaft with DC, as shown in Figure 14-3, involves using a potentiometer for controlling the polarity and magnitude of the voltage applied to each coil. Since the potentiometer makes it possible to vary the magnitude of the voltage, the magnet will take any position around the circle. While this arrangement produces accurate results, it is not satisfactory as a practical remote indicating device since it requires independently controlled potentiometers for each coil.

#### POTENTIOMETER AND MAGNET REMOTE INDICATING SYSTEM

A practical DC remote indicating system that uses the principles described in positioning shafts with DC is the Potentiometer and Magnet Indicating System as shown in the illustration in Figure 14-4. Sometimes this system is called a DC synchro.

The potentiometer in this remote indicating system is circular. DC voltage is applied to it through contacts that slide along the inner side of the potentiometer. These contacts are insulated from each other to prevent the DC voltage from being shorted out. In the three-coil permanent magnet arrangement at the right of the figure, the coils may be connected either in delta or in wye. In either connection there is little difference in the operation of the system.

The potentiometer is step-wound to compensate for the shunting effect between the potentiometer coils. This type of winding, however,

causes the voltages at points A, B, and C to vary as the potentiometer rotates in a counter-clockwise direction, as shown in Figure 14-5 in the graph of the voltages at different positions of the potentiometer contacts. All are with respect to the negative (ground) side of the DC voltage supply.

As previously mentioned, there is little difference between the operation of a wye- and a delta-connected system. Thus, if you follow through the theory of operation of the wye-connected type of three-coil-permanent magnet assembly, you will be able to understand the delta-connected type also.

During operation of the wye-connected system when the contacts are at the 0° position (+ to A and - to D), the voltages at A, B, and C respectively are 27v, 9v, and 9v. One end of coil 1 is connected to point A; one end of coil 2 is connected to B; and one end of coil 3, to C. The other ends of each coil are connected together. Current flows from B to A through coils 2 and 1, and from C to A through coils 3 and

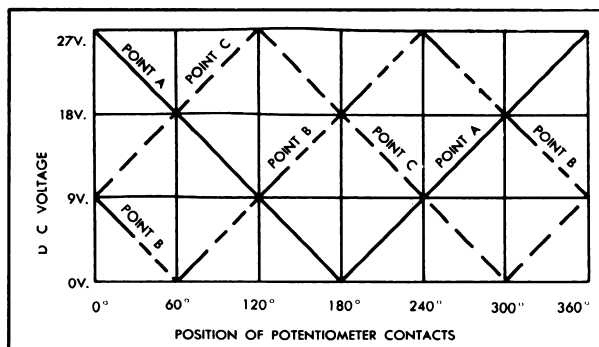


Figure 14-5. Graph of Voltage vs. Position of Contacts



1. Since B and C are at the same potential, there is no current flow between them. Coils 1, 2, and 3 become magnetized, with their poles as shown in Figure 14-4. Coil 1 carries more current and therefore exerts a strong attraction for the north pole of the magnet. Coils 2 and 3 carry equal currents and attract the south pole of the magnet with equal force. The magnet will therefore assume the position shown (the  $0^\circ$  position).

When the potentiometer contacts move to the  $60^\circ$  position, then E is at 27v, and B at 0v. As you can see in the graph in Figure 14-5, A is at 18v, and C at 18v. This causes current to flow from B to A through coils 2 and 1 and from B to C through coils 2 and 3. No current flows from A to C. In this condition coil 2 now carries more current than the other coils and strongly attracts the south pole. The current in coil 3 is reversed. Its polarity is also reversed. Thus, coils 1 and 3 attract the north pole with equal force, and the magnet assumes the  $60^\circ$  position corresponding to the potentiometer contacts.

As the potentiometer contacts move from the  $0^\circ$  position to the  $60^\circ$  position, the voltages vary uniformly as follows: A from 27v to 18v; B from 9v to 0v; and C from 9v to 18v. Hence the current through coil 1 decreases; the current through coil 2 increases; that through coil 3 decreases to zero at  $30^\circ$  and then increases in the opposite direction. The rotation of the magnet is uniform and follows the rotation of the contacts.

Since the smooth variation of voltages is upset if the coils draw much current, this system is limited to applications where the current in the coils is not appreciable when compared with the current in the potentiometer. Since the potentiometer current is not useful except for main-

taining the voltages, the system wastes considerable power. Its principal use is as an indicating device that tells the operator the angle of elevation of the antenna. It may be used to indicate the azimuth to the operator, but he usually gets this information from the CRT.

## PHYSICAL CHARACTERISTICS OF SYNCHROS

Before going into the operation of the synchro generator and motor, it is well to first consider the construction of each. Both resemble small electric motors and are identical except for several details. Each unit contains a fixed element called the *stator*, Figure 14-6, and a movable element called the *rotor*, Figure 14-7. The stator consists of a number of coils that are placed in slots around the inside of a laminated iron field structure, very much like an ordinary AC motor. These coils are divided into three groups, spaced  $120^\circ$  apart around the inside of the field. Actually the groups overlap somewhat so that the attractive force tending to pull the rotor into position is the same for all positions of the rotor. The  $120^\circ$  spacing of windings sometimes leads a person to believe that three-phase voltages and currents are used, but such is not the case. *Only single phase voltage and current are present.*

In a typical synchro generator, the rotor consists of a single coil of wire wound on a soft iron core and mounted on a shaft in a way that the axis of the coil is perpendicular to the shaft. The ends of the core are curved so that the air gap between them and the stator is small and uniform. To make friction low, the shaft is mounted on ball bearing mounts. The rotor turns inside the stator and is electrically connected through two slip rings on the shaft.

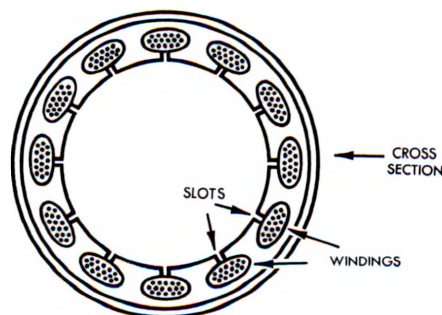
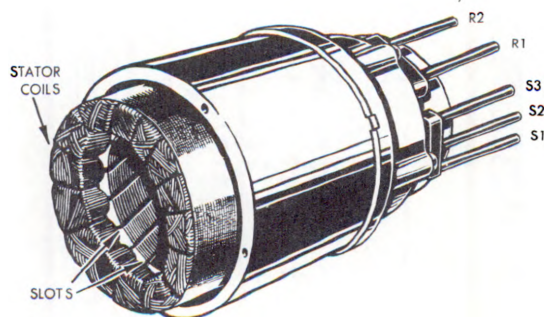


Figure 14-6. Synchro Stator



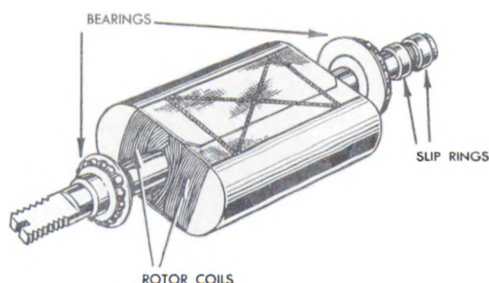


Figure 14-7. Synchro Rotor

Because a synchro motor is similar to an ordinary AC motor, it tends either to oscillate violently or to spin continuously under some conditions. This is particularly likely to happen when the shaft is turned suddenly as, for example, when power is first applied to the system. To prevent this undesirable oscillation a heavy metal flywheel, called an *inertia damper*, is mounted on one end of the shaft, Figure 14-8. This flywheel is mounted so that it turns freely on the shaft for  $45^\circ$  or so, and then runs into a

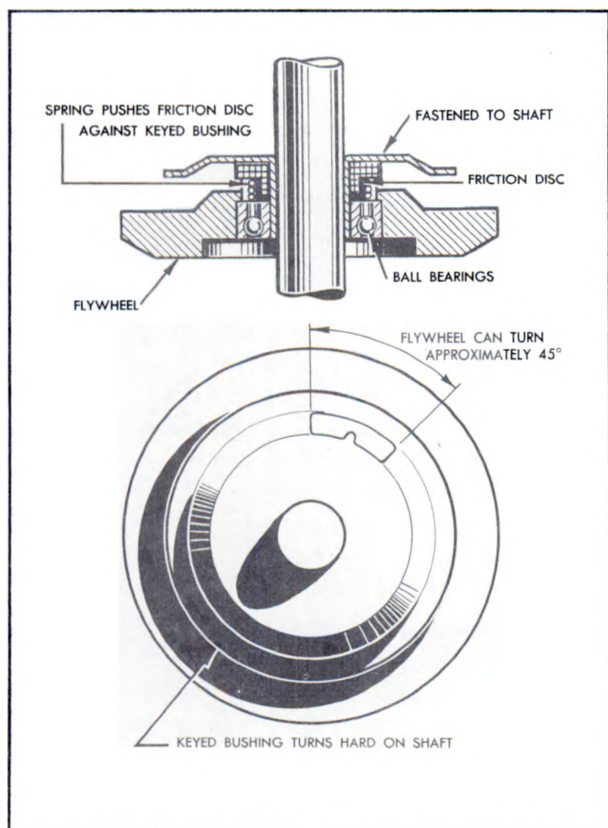


Figure 14-8. Typical Inertia Damper

keyed bushing. This bushing, which is fastened to the shaft through a friction disc, also turns on the shaft but with a great deal of friction.

For slow changes in position of the shaft, the flywheel follows along without much oscillatory effect. But if the shaft tries to turn suddenly, the flywheel tends to stand still, and the friction disc acts as a brake to slow down the motion of the shaft. This keeps the shaft from reaching a speed fast enough to start oscillating or spinning. If oscillation or spinning does occur, it is usually a certain indication that something is wrong with the damper.

In the synchro the leads to the rotor and stator are brought out through two insulating strips at the back of the motor or generator. The stator connections are marked S1, S2, and S3; the rotor connections R1 and R2, as shown in Figure 14-6. This system of marking connections is standard among the various manufacturers.

## OPERATION OF SYNCHRO SYSTEM

### Positioning a Shaft with AC Voltages

Figure 14-9 shows a solenoid located close to a permanent magnet that is mounted so that it can rotate on an axis. When a 60 cps voltage

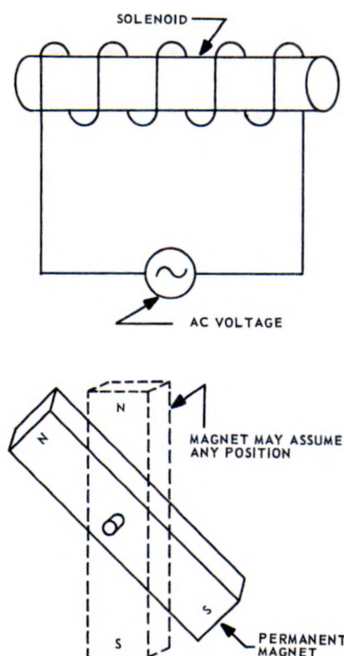


Figure 14-9. Effect of AC Solenoid on Magnet

is applied to the solenoid, the torque exerted on the magnet will reverse direction 120 times per second, first in one direction, then in the opposite direction. Since the magnet cannot reverse direction this rapidly, it will respond only to the average torque and since the average torque is zero, the magnet will assume some position independent of action by the coil. Thus, since the magnet is not affected by the coil, it is obvious that it cannot be used with an AC coil as it can be with a DC coil.

But when you substitute another AC coil for the permanent magnet, connected as in Figure 14-10, the current in both solenoids reverses at the same instant. This causes the same end of the rotating solenoid to be attracted to the nearest pole of the fixed solenoid throughout the entire cycle. As shown, the C-end of the rotating solenoid is attracted to the B-end of the fixed solenoid throughout the cycle. Reversing the connections on either coil (but not on both coils), reverses the rotating coil and causes B to attract D. This reversal of connections is equivalent to a reversal of phase in that coil.

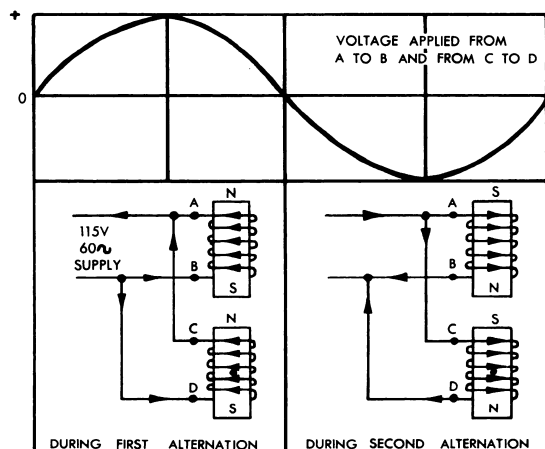


Figure 14-10. Effect of One AC Coil on Another

### Symbols for Synchro Diagrams

It is not very practical to show synchro circuits in the form of photographs or drawings of the actual apparatus, since they merely show the outer appearance and leave the inside obscure. Therefore, to make it easier to describe the operation of a synchro motor (or generator), standard diagrams which are referred to as

*synchro diagrams* (Figure 14-11) are used. In forming these diagrams, there is a standard way for describing the shaft position. This is necessary to insure that if the rotor shaft is properly connected mechanically in the system, it will turn to the correct position when it is connected electrically to a synchro generator.

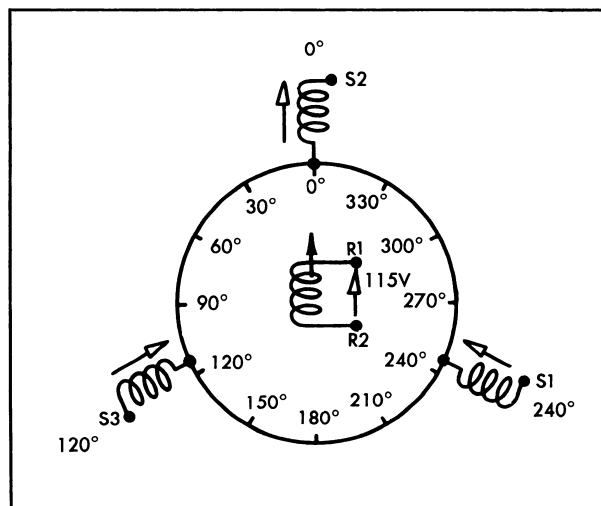


Figure 14-11. Synchro Diagram

In Figure 14-12 the position the rotor takes when it lines up with the stator coil that is connected to S2 is called the *electrical zero position*. Other positions are measured in degrees in a counterclockwise direction (looking at the shaft end of the unit).

In studying the synchro diagram think of the rotor as turning around an axis in the center of the diagram, and that the arrow on the R1 end of the rotor points at the number indicating the electrical position of the rotor.

Another point to consider is that the rotor coil in the diagram is assumed to be wound in the same direction, going from R1 to R2, and that each of the stator coils is wound from the outside to the inside.

The open-headed arrows are important in that they indicate the following *relative phase* polarity. If two arrows point in the same direction, the voltages are in phase; if the arrows point in opposite directions, they are 180° out of phase.

All voltages are either in phase or 180° out of phase with each other. The diagram indicates that the voltages from R2 to R1, from

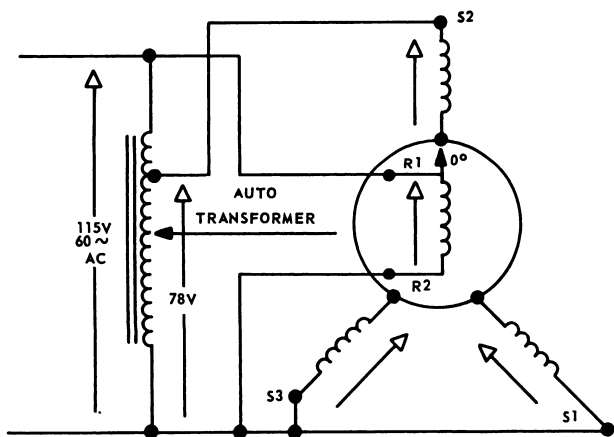


Figure 14-12. 0° Position

common to S2, from S1 to common, and from S3 to common are in phase, while from S2 to common they are 180° out of phase with either S1 to common or S3 to common.

#### Voltages Required to Position a Synchro Motor

For a standard synchro motor to turn to the electrical zero position, 115v AC (rms value) must be applied to the rotor leads, zero volts between S1 and S3, 78v between S2 and S1-S3 in such a way that the voltage from R1 to R2

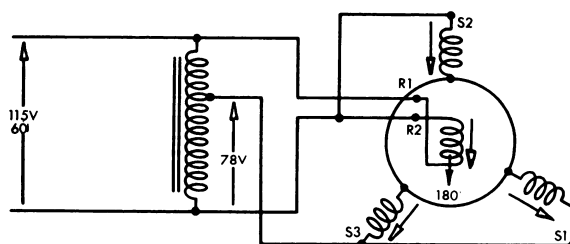


Figure 14-13. 180° Position

is as shown in Figure 14-12. In this condition the coils connected to S1 and S3 exert equal attractive forces on the R2 end of the rotor. The coil connected to S2 attracts the R1 end of the rotor with greater force, for it carries the current of both the other coils. Therefore this causes the rotor to assume the 0° position.

Reversing the leads to the 78v supply causes the voltages across the stator coils to reverse in phase and the rotor coil to reverse in position, causing it to point to the 180° position, Figure 14-13.

Connecting together the other pairs of stator leads and applying the 78v to the third lead either in phase or out of phase with the R1 or the R2 voltage will cause the rotor to turn to the four positions — 60°, 120°, 240°, 300° — as shown in Figure 14-14.

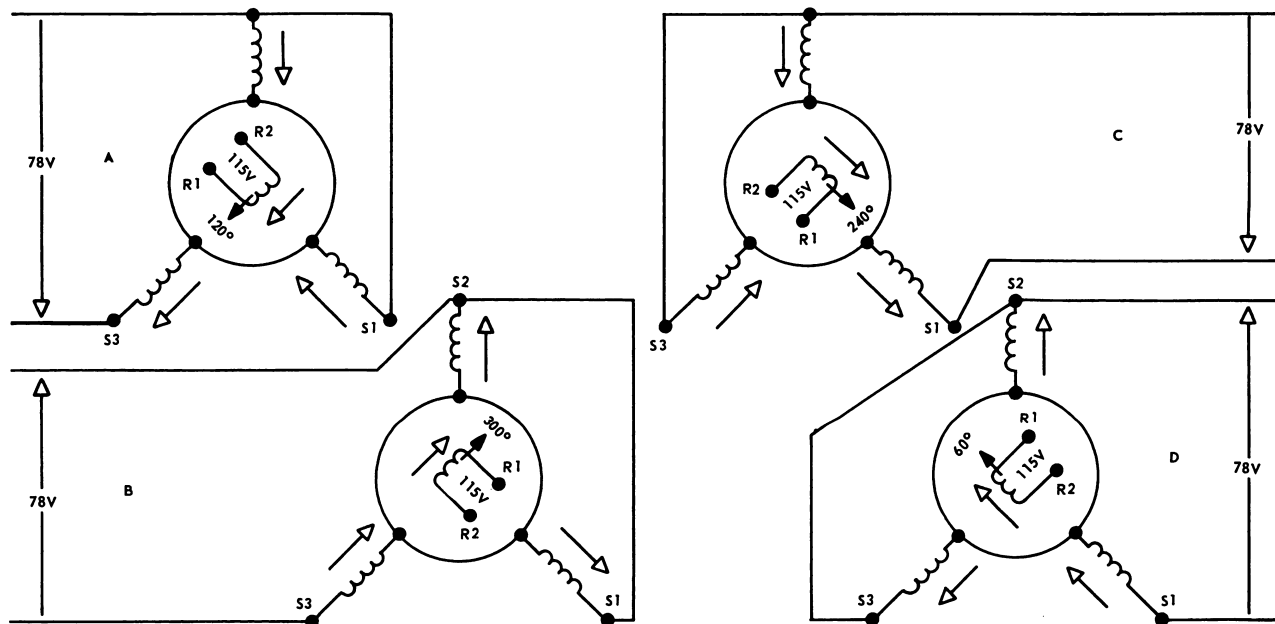


Figure 14-14. Four Positions Possible With Two Leads Connected

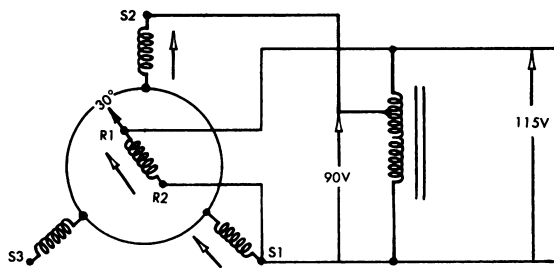


Figure 14-15. 30° Position

When S3 is left open and 90v is applied between S1 and S2 (90v is required to produce the same attraction as before), the rotor will come to rest in the 30° position. This results because the R1 end of the rotor is attracted to the coil connected to S2 with the same force as the R2 end is attracted to the S1 coil, as you can see in Figure 14-15. Reversing the phase polarity of the 90v causes the rotor to turn to the 210° position (top of Figure 14-16). By leaving S1 or S2 open and applying 90v to the remaining two terminals, either in phase or 180° out of phase, you can produce the four other positions shown in Figure 14-16.

So far, only the means of obtaining 12 positions, spaced 30° apart, has been described. Next, consider methods for obtaining positions at intermediate points between these positions.

Figure 14-17 shows a setup that gives positioning at any point from 0° to 60°. In it, fixed AC voltages are applied to S1 and S2. The voltage at S3 is variable from the value of the voltage at S1 to that at S2 with a variable autotransformer. If the voltage at S3 is the same as that at S2, the rotor will turn to the 60° position as shown in Figure 14-14D. If the voltage at S3 is equal to the voltage at S1, the rotor turns to the 0° position as indicated in Figure 14-12. These two conditions occur when the variable tap is at N and P respectively. When the tap is at M, midway between N and P, S3 is at zero potential with respect to the common connection of the coils and exerts no attraction or repulsion. Hence, the rotor turns to the 30° position just as it did when S3 was disconnected.

At intermediate points between M and N, the phase of the voltage in coil S3 will be as indicated by the dotted phase arrow and the rotor will assume a position between 30° and 60°, depending upon the magnitude of the voltage at

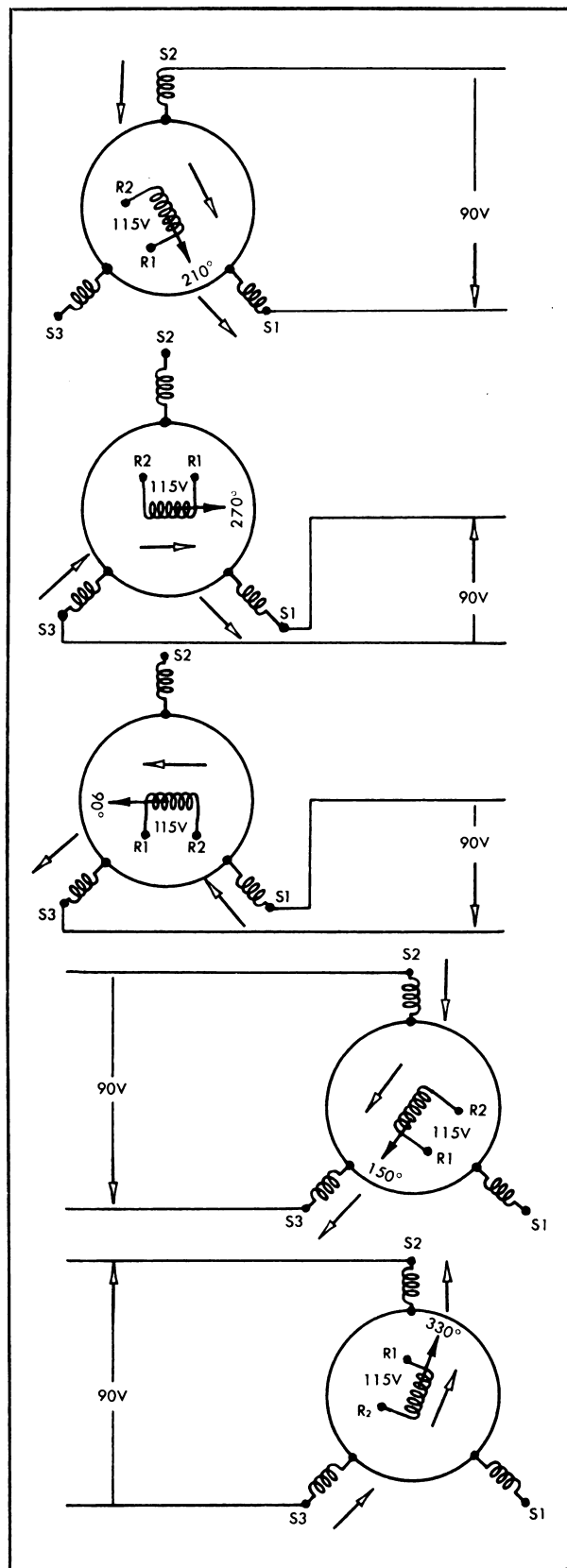


Figure 14-16. Other Positions with One Lead Open

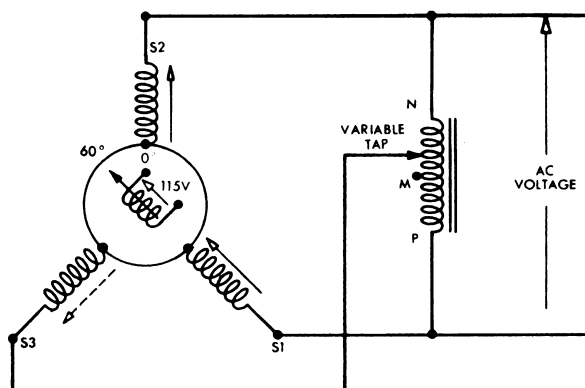


**Figure 14-17. Using Variable Tap to Obtain Positions from 0° to 60°**

S3. Between N and P the rotor turns to some point between 30° and 0° which depends on the magnitude of the voltage at S3, the phase polarity of which is opposite to that indicated by the arrow.

By using stator voltages intermediate to the voltages for the 12 basic positions, it is possible for the rotor to assume any position around the circle. The voltages necessary to produce any angular position may be plotted as in the graph in Figure 14-18. The three curves show the voltages between pairs of stator leads. The voltages above the zero axis are in phase with the voltage from R1 to R2, and those below the axis are opposite in phase. Keep in mind that the curve of the S2 to S1 voltages will be 180° out of phase with the curve shown for the S1 to S2 voltage.

Due to the method of arranging the rotor and stator windings on a standard synchro, the curves shown in the graph in Figure 14-18 are in the shape of sine curves. In other words, the curves look like time graphs of sinusoidal voltages. They show something entirely different, however, for time is not involved in any way in the preceding graph. Another thing to remember is that while they have the appearance of the voltages in three phase, they are really single phase. What is actually shown in the curves is the effective values of the three AC voltages plotted against position of the rotor, and whether the voltages are in phase or 180° out of phase with the rotor voltage.

Next, suppose you check the voltages in Figure 14-18 against the voltages in some of the cases previously discussed. Notice that when the rotor

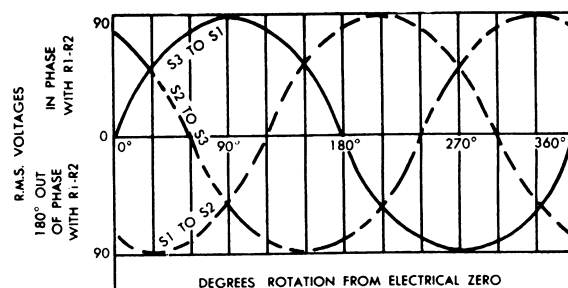


**Figure 14-18. Voltage vs. Rotor Positions**

turns to the 0° position, the voltage from S3 to S1 is 0v, and the other two are 78v. This is the same situation shown in the 0° position illustrated in Figure 14-12. When the voltage between S1 and S2 is increased to 90v, each of the others is 45v; the shaft will turn to the 30° position as in Figure 14-15. In Figure 14-15, S3 is open and the 90 volts from S2 to S1 divide equally between the coils, placing 45 volts across each coil. Again using voltages from Figure 14-18, when S3 to S1 and S1 to S2 is 78 volts, you have the same situation as in Figure 14-14D; the shaft turns to 60°. At each position of the rotor there is a certain definite value and phase condition for each of the three voltages.

### Transformer Action of Synchro

Another important factor in synchro operation is transformer action. This deals with the voltages induced in the stator windings when an AC voltage is applied at various positions of the rotor. In studying this action, assume that at first the stator leads are open. Figure 14-19 shows one stator coil with the rotor in three positions. At A the axis of the rotor and stator are aligned, and the flux produced in the rotor due to the 115v applied to it induces a voltage in the stator coil. Due to the turns ratio this induced voltage in a standard synchro is 52 volts. This voltage is in phase with the rotor voltage as indicated by the open-headed arrows.

In B the rotor is in the 90° position. In this position the flux set up in the rotor is not effective in inducing a voltage in the stator because the flux that passes through the coil is at right angles to its axis and induces equal and opposite voltages on the opposite sides of each turn. Thus, at this position the net voltage is zero.

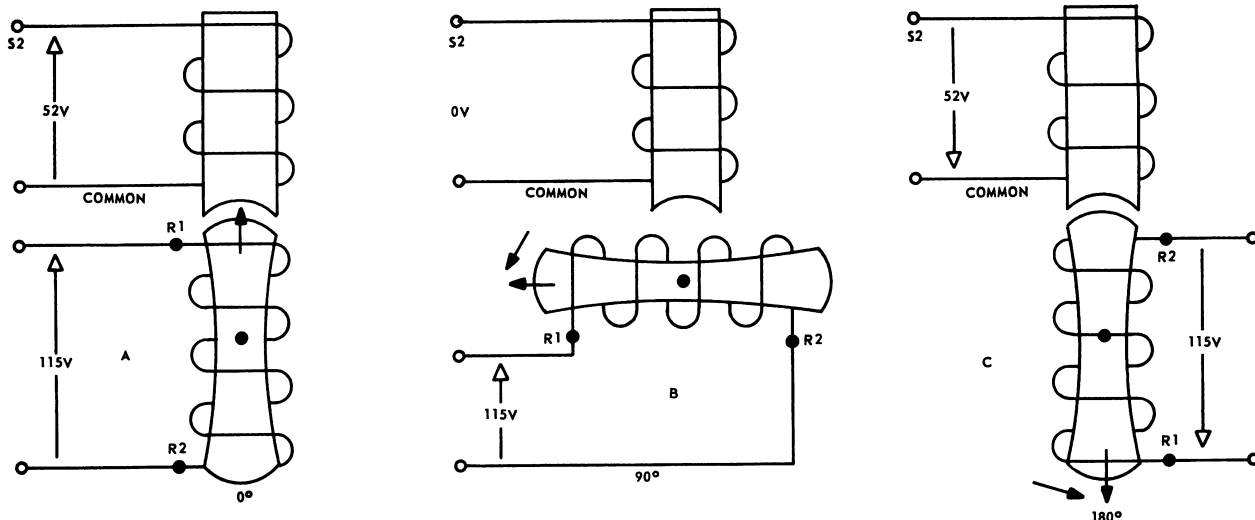


Figure 14-19. Transformer Action

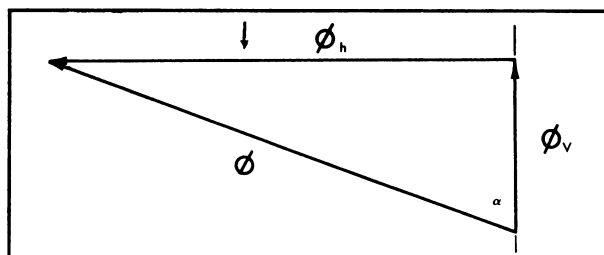


Figure 14-20. Components of Flux  $\phi$

In C the rotor is in the  $180^\circ$  position. Here the flux at any instant is  $180^\circ$  out of phase with what it would be in the  $0^\circ$  position. Thus, the voltage induced in the stator coil is  $180^\circ$  out of phase with the rotor voltage (R1 to R2). The magnitude is the same as for the  $0^\circ$  position, 52 volts.

At intermediate points the magnitude and relative phase polarity depend upon the flux threading the coil. If you resolve the flux ( $\phi$ ) into the components  $\phi_h$  and  $\phi_v$ , as shown in Figure 14-20, you can determine the relation between the flux and the induced voltage. The horizontal component is ineffective, while the vertical component of flux induces a voltage in the coil. But  $\phi_v/\phi$  is the cosine of angle  $\alpha$  between the axis of the stator and rotor. Thus, you can see that for any angle of rotation the voltage induced in the stator coil S2 is  $52 \cos \alpha$  volts. (Figure 14-21 is a graph of these voltages vs. shaft position.) Due to the  $120^\circ$  spacing of the

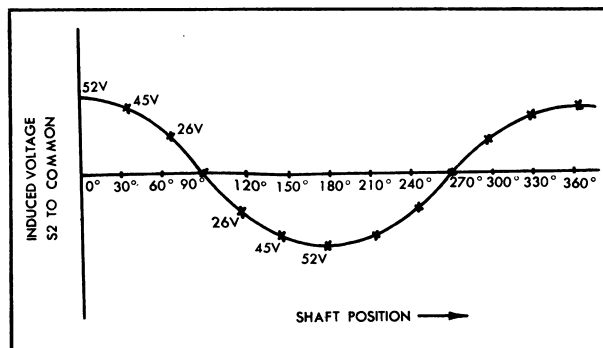


Figure 14-21. Induced Voltage vs. Shaft Position

stator coils, maximum voltage is induced in S3 at the  $120^\circ$  position and in S1 at the  $240^\circ$  position.

Since the connections are made to points S1, S2 and S3, the voltages appearing between those points are of special interest. These voltages are the sums of, or differences between, the voltages induced in the individual stator coils, depending on whether the voltages are in phase or  $180^\circ$  out of phase. For example, the voltage from S1 to S2 is the sum of the voltages from S1 to the common point and from the common point to S2. Figure 14-22 shows the magnitude of the voltages between stator connections for any position of the rotor. In comparing the voltages shown in this figure and those shown in Figure 14-18, note that the voltages induced in the stator coils



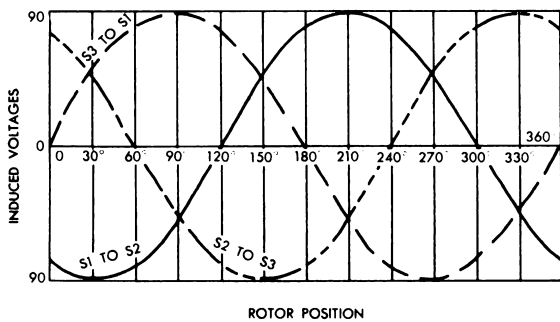


Figure 14-22. Induced Voltage vs. Rotor Position

are the same as the voltages required to position the rotor to the same angle.

### How a Synchro Motor Follows a Generator

The synchro system in its simplest form consists of one motor connected to one generator as shown in Figure 14-23. With this connection, whenever the shaft of the generator turns, the shaft of the motor turns. The *electrical position* of the motor shaft is the same as that of the generator shaft. Thus, for example, when the generator shaft turns to electrical zero, the motor shaft turns to  $0^\circ$ , or when the generator shaft turns to  $30^\circ$ , the motor shaft turns to  $30^\circ$ , and so on. In order to see how the motor shaft follows the generator shaft, first examine the internal conditions of each synchro. To do this, consider both shafts at the same position and then both shafts at different positions.

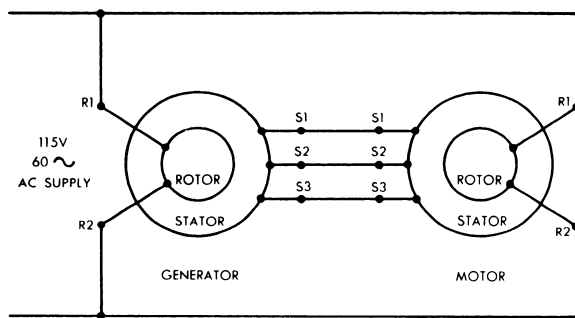


Figure 14-23. Synchro System Connections

Figure 14-24 shows both shafts turned to the  $0^\circ$  position. In this position there will be voltages induced in the stator windings of both generator and motor, with the phase polarity and

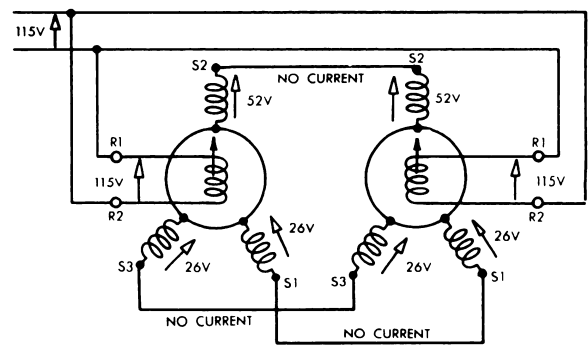


Figure 14-24. Generator and Motor Both at  $0^\circ$  Position

magnitude indicated. Note that the voltages of the motor are equal to those of the generator, but because of the method of connection, the voltages in one oppose the voltages in the other. This causes no current to flow in the stator coils. Hence, there is no magnetic field set up by them and thus there is no attraction or repulsion to cause either rotor to turn. The rotors will remain in the  $0^\circ$  position indefinitely if no change in outside conditions occurs. The same situation exists if the two rotors are at any other position, except that the magnitudes of the induced voltages vary in accordance with the curves shown in the induced voltage versus rotor position graph in Figure 14-22.

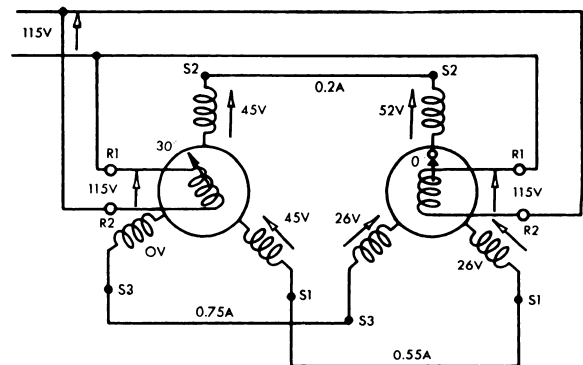


Figure 14-25. Shafts in Different Positions

Figure 14-25 shows generator and motor shafts at different positions. When the two shafts occupy different positions, the voltages induced in the stator windings are not balanced as they are when the shafts occupy the same position. Suppose, for example, that the generator rotor

is at the  $30^\circ$  position and that the motor rotor is at the  $0^\circ$  position. By referring to Figure 14-22 showing rotor position, you can determine the value of the induced voltages. Note that there is no balance among the voltages (Figure 14-25) and that currents flow in all three stator leads and all stator coils of both generator and motor.

The currents are greatest in the circuits where the voltage unbalance is greatest, that is, in the circuit composed of S3 coils and common. The effect of these currents is to produce magnetic fields that tend to turn the motor shaft to the same position as the generator shaft. Thus, if the motor shaft is free to turn it will take the  $30^\circ$  position. As the motor shaft turns, the misalignment decreases and, in turn, the unbalance of voltages decreases and the currents produced by them decrease. When the shaft reaches the  $30^\circ$  position, the same conditions that prevailed when both shafts occupied the same position will prevail.

A torque also is set up in the generator which causes it to turn in the clockwise direction, resulting in both rotors tending to align themselves. In the usual system the generator shaft is attached to some device so that the shaft is not free to rotate.

To see more clearly the magnetic effects producing the torques influencing motor operation, note in Figure 14-26 the conditions in the circuit at a given instant. The same conditions do not hold except for that particular instant. Examination of the figure shows the following action: (1) the lower end of the S2 coil of the generator attracts the R1 end of the rotor, (2) the S end of the S1 coil repels the R2 end of the rotor, and

(3) the N end of the S3 coil attracts the R2 end of the rotor. All these forces tend to cause rotation in the clockwise direction.

In the motor the magnetic poles of the stator coils are the reverse of those of the generator. Consequently, the forces of attraction and repulsion are reversed and the torque is counter-clockwise. At one-half cycle (of the input voltage) later, the polarities of the magnets and the directions of current flow will reverse. The forces of attraction and repulsion do not change, however, since the polarities of the stator and rotor magnets reverse at the same instant.

### How Currents Flowing in a Synchro System Depend upon Shaft Positions

As was shown, current flows in the stator circuits whenever the two shafts occupy different positions. In any stator lead the amount of this current flow depends on the difference between the voltage induced in the two coils to which that lead connects. Suppose that the current in any one stator lead (S2 for example) is measured and that the actual shaft positions are adjusted so that this current is maximum for each difference of the shaft positions. From Figure 14-27, which shows this maximum possible current versus the difference in shaft positions, you see that currents as high as 3 amperes are possible if the shafts are held in positions differing by  $180^\circ$ .

But in practical operation the shafts are seldom more than a degree apart, since the motor shaft starts turning as soon as the generator shaft starts to rotate. Hence, the current in any lead seldom becomes more than one-tenth of an ampere. The current in one stator lead depends not only on the difference of shaft positions, but

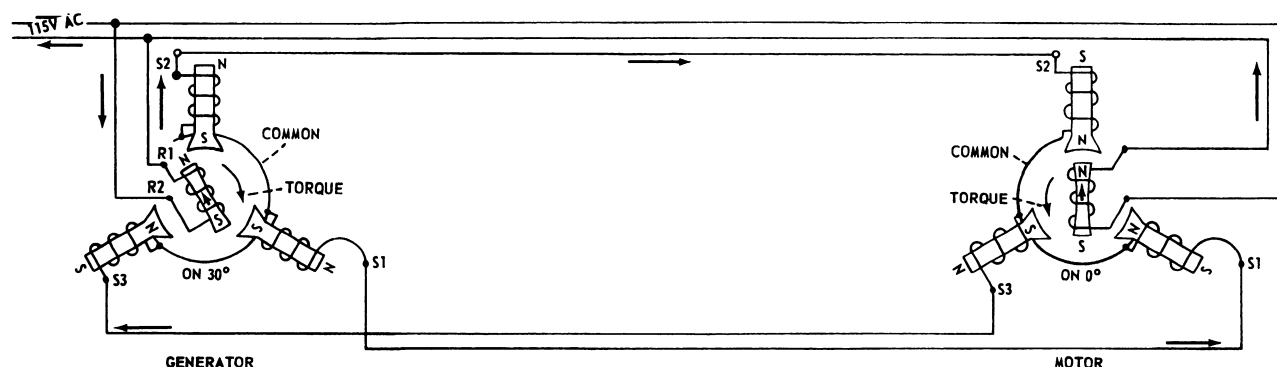


Figure 14-26. Instantaneous Forces, Shafts in Different Positions

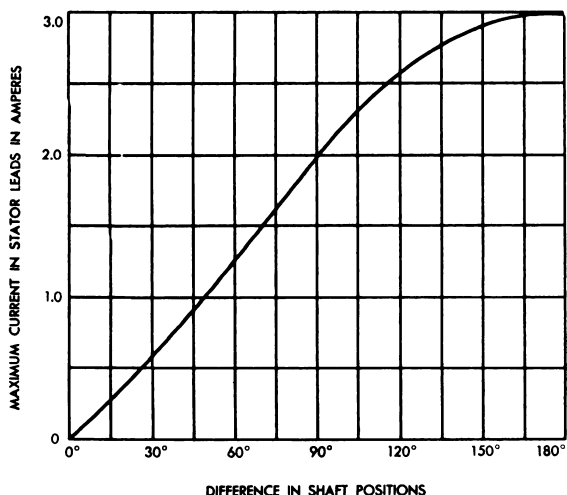


Figure 14-27. Stator Current vs. Difference in Shaft Positions

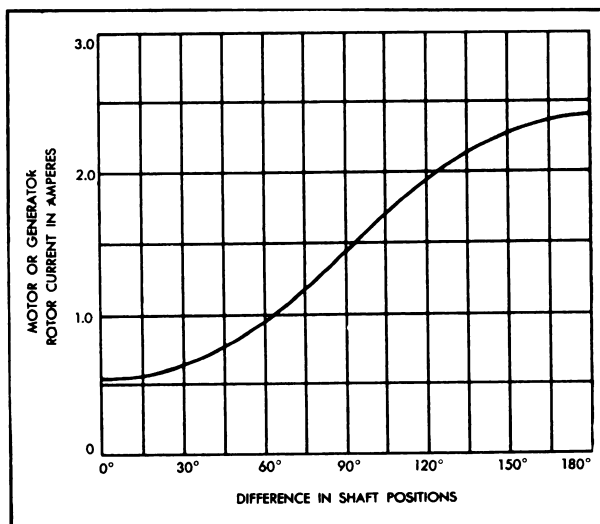


Figure 14-29. Rotor Current vs. Difference in Shaft Positions

also on the positions themselves. Figures 14-28 and 14-29 show the percent of maximum current in each lead for the various mean shaft positions (average positions between the two shaft positions).

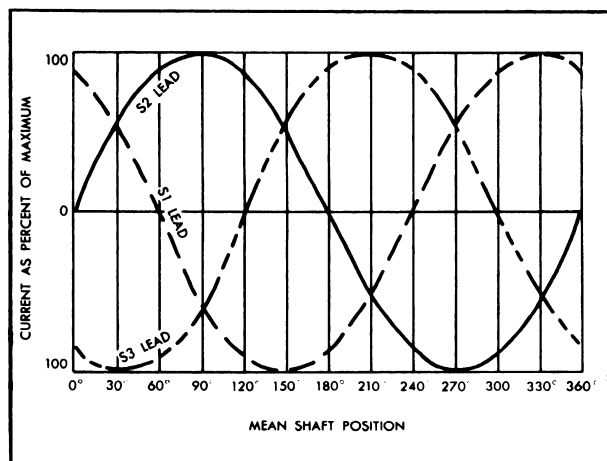


Figure 14-28. Stator Current vs. Mean Shaft Position

To illustrate how these two figures may be used in conjunction, take the case where the generator shaft is at 60° and the motor shaft is at 120°, a difference of 60°. By referring to the difference in shaft position diagram, you can see that the maximum possible current in any one lead is 0.95 ampere. By referring to the mean

shaft position chart you see what percentage of this maximum current flows in each lead with a mean shaft position of 90°. The S2 lead carries 100% or 0.95 ampere, and the S1 and S3 leads carry 50% or 0.475 ampere. The fact that the current S1 and S3 carry is shown below the zero axis means that the phase is opposite that of the current flowing in the S2 lead. Remember that currents are usually not high since small angular differences are the usual or normal condition.

Since the synchro generator or motor acts like a transformer, an increase in stator current causes a corresponding increase in rotor current. (See Figure 14-29). Although all three stator currents are zero when the shafts are aligned, the rotor current is still approximately 0.5 ampere. As in any transformer the primary draws some current with no load on the secondary. This current produces magnetization of the rotor and supplies its losses.

The following statements summarize the factors concerning the relative shaft positions:

- Reversing the rotor leads on either motor or generator causes a 180° difference in shaft positions.
- Interchanging the S1 and S3 leads on either the generator or the motor causes rotation of the motor shaft in the opposite direction to that of the generator.
- Shifting the stator leads so that S1, S2, and S3 of the generator connect to S2, S3, and S1 of

the motor produces a fixed difference of  $120^\circ$  in shaft positions.

Since both the synchro motor and the load attached to it have some friction and inertia, some torque is required both during motion and during acceleration. Since no torque is developed when the two rotor positions coincide, perfect accuracy is impossible. Some angular displacement is necessary to overcome friction and inertia. This lag is not appreciable; in most cases it is only a fraction of one degree. To minimize the inaccuracy, the friction of the motor must be as low as possible and the torque gradient (torque per degree displacement) must be as high as practicable. The torque gradient depends upon the internal impedance of its stator coils. Hence, both the motor and generator must be large enough, for if either is too small, the higher impedance of the windings will reduce the current for small unbalances of voltages and the torque will accordingly be less.

Note in the graph of the torque versus angular displacement relations for a typical synchro (Figure 14-30), that for small angles the graph is practically a straight line. The term *torque gradient* is simply the slope of the curve in that region. Maximum torque is developed at approximately  $100^\circ$ . Figure 14-30 shows zero

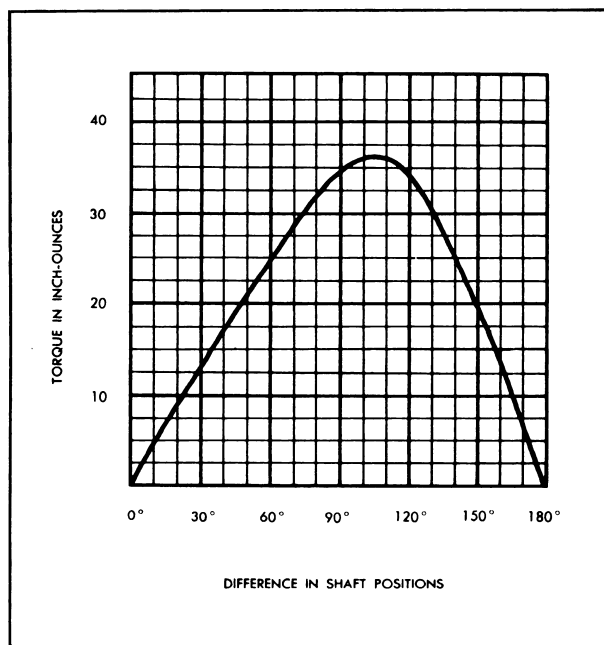


Figure 14-30. Torque vs. Displacement

torque at  $180^\circ$  displacement. This is true because the powerful forces that are exerted are in a state of balance. This is, therefore, an unstable condition and any small change in either rotor position results in a return to the zero displacement condition.

One synchro generator is sometimes used to transmit data to several motors at various locations. For example, the synchro generator connected to the gyrocompass aboard a plane is used for driving synchro motors at several of the compass repeater stations. If there are more than two motors, a larger generator must be used to maintain accuracy. If any synchro motor becomes jammed, it affects the accuracy of the entire system since it acts as a generator competing with the real synchro generator. As shown in Figure 14-31 all the corresponding stator and rotor coils are connected in parallel.

### THE DIFFERENTIAL SYNCHRO

In a mechanical system a differential connects three shafts together in such a way that the amount that one shaft turns is equal to the difference between the amounts that the other two turn. The differential synchro is named according to the two functions it performs. If the shaft of the differential synchro serves to indicate the difference in shaft positions of two other synchros, the other two synchros are generators and the differential synchro is a *motor*. If the differential synchro shaft position is to be subtracted from that of a synchro generator (or vice versa) and the difference to be indicated by the rotor position of a motor, the differential synchro is then a *generator*.

The stator of the differential generator or motor is very similar to the stator of the ordinary synchro. It consists of three sets of coils wound in slots and spaced  $120^\circ$  apart around the inside of the field structure. The rotor, however, differs considerably from that of an ordinary synchro. It is cylindrical in shape and has *three* sets of coils wound in slots and equally spaced around the circumference, Figure 14-32. Connections to the external circuits are made through three brushes riding on three slip rings on the rotor shaft. As in the ordinary synchro, the motor has an oscillation damper and the generator does not. There is no connection to the 110v AC as in the case of

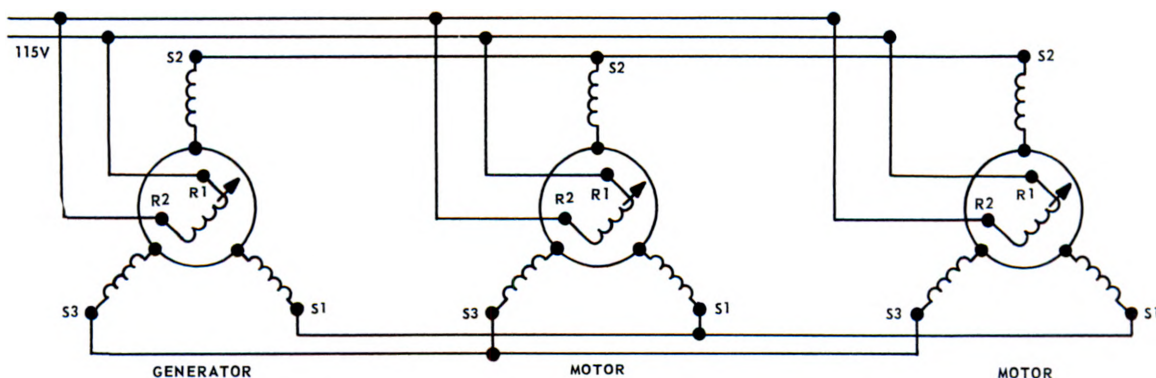


Figure 14-31. Two Motors Connected to One Generator

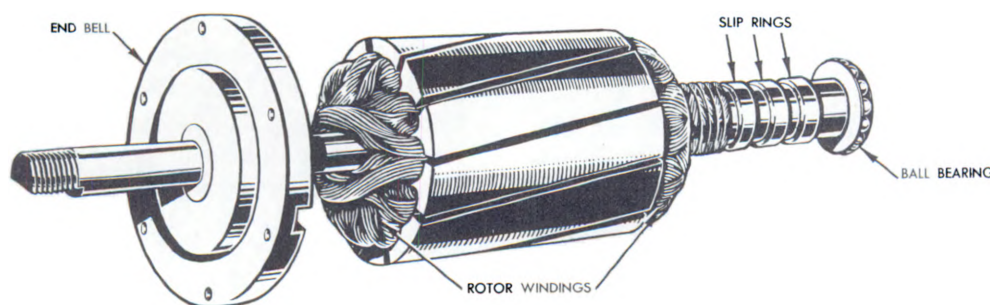


Figure 14-32. Rotor of Differential Synchro

the ordinary synchro. The three leads to the rotor are designated R1, R2, and R3. Electrical zero is the position in which rotor coil R2 is aligned with stator coil S2, R1 with S1, and R3 with S3.

Electrically, the differential synchro acts as a 1:1 transformer with the stator coils as the primary and the rotor coils as the secondary. Due to the air gap, there must be a few more turns on the stator coils than on the rotor; hence, the transformation is not a 1:1 ratio when the rotor coils are connected as the primary.

The synchro generator in Figure 14-33 is connected to a differential synchro in which the rotor leads are open. Both shafts are in the  $0^\circ$  position. Since the stator coils of the differential synchro connect to the stator coils of the synchro generator, the voltages are equal in magnitude and are in phase. Since the differential is a 1:1 transformer, the induced voltages are equal in magnitude and are in phase with the stator voltages. When the generator shaft is turned to some other position, the voltages in the stator coils of the differential and, consequently, those induced in

the rotor coils, will be the same as the voltages in the stator coils of the generator.

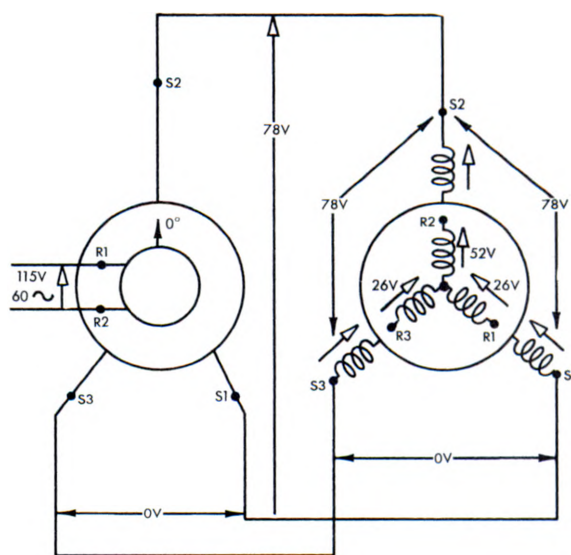


Figure 14-33. Differential Transformer Action at  $0^\circ$  Position

As the generator shaft is rotated with the differential shaft remaining in the  $0^\circ$  position, the rotor voltages of the differential will be as shown in the rotor position diagram in Figure 14-24. A motor with its stator windings connected to the rotor windings of the differential will have the correct voltages applied causing it to assume the same position as the generator rotor. This means that with the differential rotor in the  $0^\circ$  position, the system functions as it would if the differential were out of the circuit and connections were directly from generator to motor.

When the generator rotor is in the  $0^\circ$  position, and the rotor of the differential is turned to  $15^\circ$  as shown in Figure 14-34, the voltages induced in the rotor coils will be as indicated. The voltages between terminals R1-R2, R2-R3 and R3-R1 are respectively 63v, 87v, and 24v, and the R2-R3 voltage is in phase with the R1-R2 voltage across the rotor of the generator. The other voltages are  $180^\circ$  out of phase. These voltages correspond to the voltages induced in the stator windings of a synchro generator in which the rotor is at the  $345^\circ$  position. To understand this action, consider the relative angle of the inducing magnetic field axis and the axis of the secondary coils.

When a synchro generator rotor is in the  $345^\circ$  position, the field axis is  $15^\circ$  clockwise from the axis of the S2 secondary coil; the voltage induced

in it equals  $52 \cos (-15^\circ)$ . Likewise the voltages induced in S1 and S3 equal  $52 \cos (-255^\circ)$  and  $52 \cos (-135^\circ)$ . These voltages set up a field in the motor that equals the vector sum of the three fields of the stator coils, resulting in a field with an axis at the  $345^\circ$  position. The rotor of a synchro motor accordingly moves to the  $345^\circ$  position. Now in the differential synchro, the composite primary field has reached the  $0^\circ$  position while the secondary (rotor) coils have rotated  $15^\circ$ . This makes the field axis  $15^\circ$  clockwise from the axis of the R2 coil.

The induced voltages then have the values previously stated and shown in the transformer action illustrated in Figure 14-34. If the motor had its stator coils connected to the terminals R1, R2, R3 of the differential, the voltages applied to it would cause the rotor of the motor to move to the  $345^\circ$  position. Conversely, if the rotor of the differential were turned to the  $345^\circ$  position, the motor would turn to the  $15^\circ$  position. If the generator shaft position is at  $0^\circ$ , the voltages induced in the rotor of the differential synchro are the same as if the rotor of the differential were at  $0^\circ$  and the rotor of the generator were rotated in the opposite direction.

Thus, whenever the shafts of both generator and differential rotate, the resultant field and induced voltages vary with the difference of the angular position of the shafts. A motor connected to the rotor coils of the differential will assume a shaft position equal to the difference of the angular positions of the other two shafts. The mathematical proof, quite a long one, is beyond the scope of this manual. Briefly, it involves setting up equations for the voltages induced in each of the rotor coils of the differential by each of the stator coils and adding the induced voltages in pairs.

A qualitative explanation of what causes the motor shaft to turn to the indicated position is fairly simple to understand. For example, take the case illustrated in Figure 14-35. The generator rotor is at the  $180^\circ$  position. Hence, the voltages induced in S1, S2, and S3 are 26v, 52v, and 26v, respectively, in the phases indicated. All the voltages in S1, S2, and S3 of the differential are identical. These voltages set up a magnetic field in the differential that has the same position,  $180^\circ$ , for its axis as the field in the generator. If the axis of the rotor of the differential generator

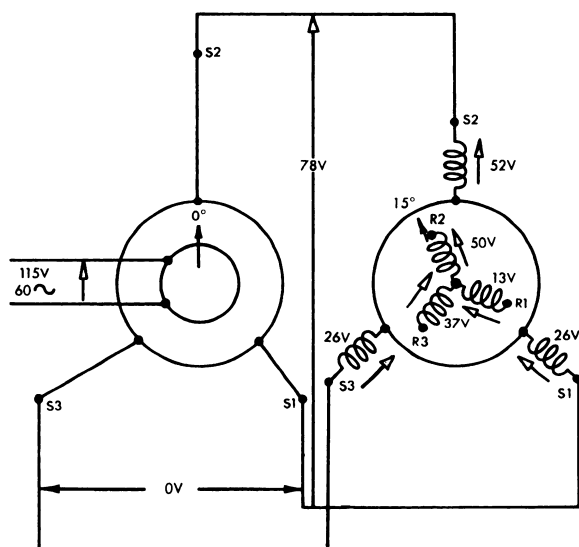


Figure 14-34. Transformer Action at  $15^\circ$  Position



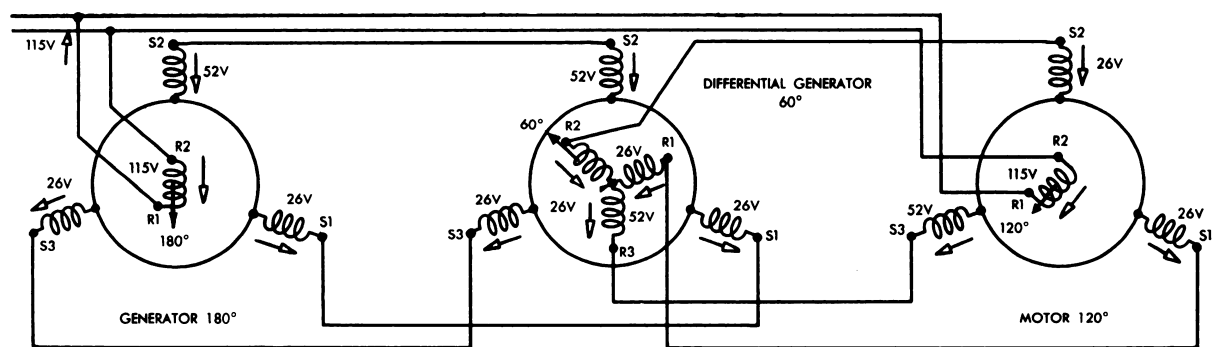


Figure 14-35. Synchro System with Differential Generator

were aligned with the  $0^\circ$  position, the induced voltages would be identical to the stator voltages. But since the rotor coil axis is shifted to the  $60^\circ$  position, the axis is only  $120^\circ$  from that of the magnetic field.

Therefore, the voltages induced in R1, R2, and R3 are  $52 \cos 120^\circ$ ,  $52 \cos 240^\circ$ , and  $52 \cos 0^\circ$  volts, respectively. The voltages from R1 to R2, from R2 to R3 and from R3 to R1 are therefore 0v, 78v, and 78v with R2-R3  $180^\circ$  out of phase with R1-R2 of the generator. When these voltages are applied to the motor, the rotor will turn to such a position that the voltages induced in its stator coils by the rotor just balance these voltages. This happens when the rotor is at the  $120^\circ$  position, the difference between  $180^\circ$  and  $60^\circ$ .

By reversing pairs of leads either between the generator and the differential or between the differential and the motor, you can make any one of the shafts assume a position equal to the sum or difference of the angular positions of the other shafts. The proper connections for various types of operation are shown in Figure 14-36.

A differential synchro cannot be a 1:1 transformer in both directions. Therefore, the perfect balance of voltages indicated is not strictly accurate. The differential synchro is so designed that the voltages induced in the rotor coils are equal to the voltages applied to the stator windings provided the shaft is at  $0^\circ$ . When a differential is connected between a generator and a motor, some current flows in the generator's stator circuit, reducing the output voltages due to the IR drop. These lower than normal voltages are induced in the rotor windings of the differential. They are too small to buck out the motor voltages, and some current flows in the motor stator

circuits. The generator currents, usually higher, reduce the accuracy of the system. Therefore, when a very high degree of accuracy is required, the use of a differential is avoided if at all possible.

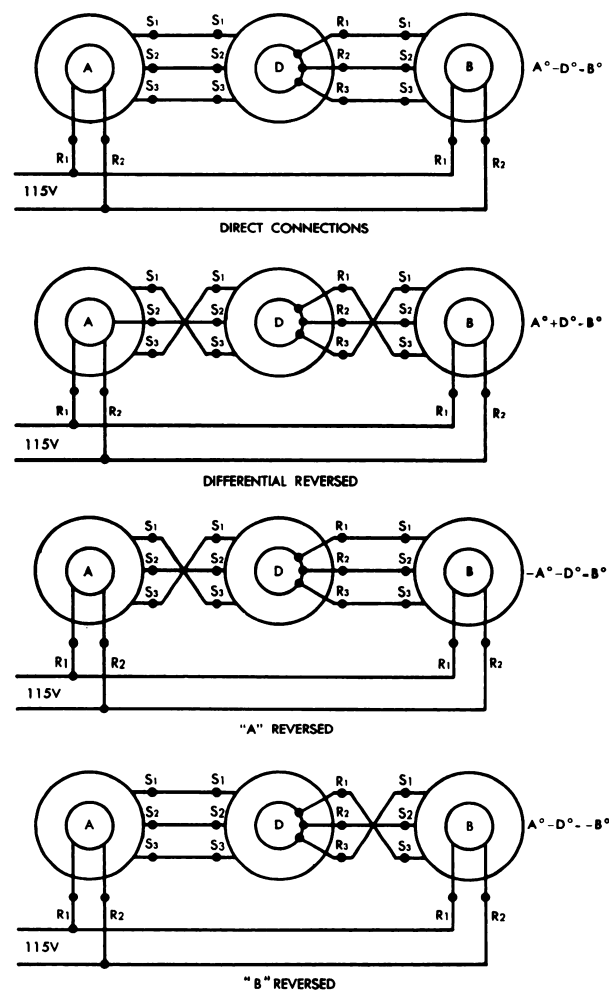


Figure 14-36. Various Connections for Synchro System

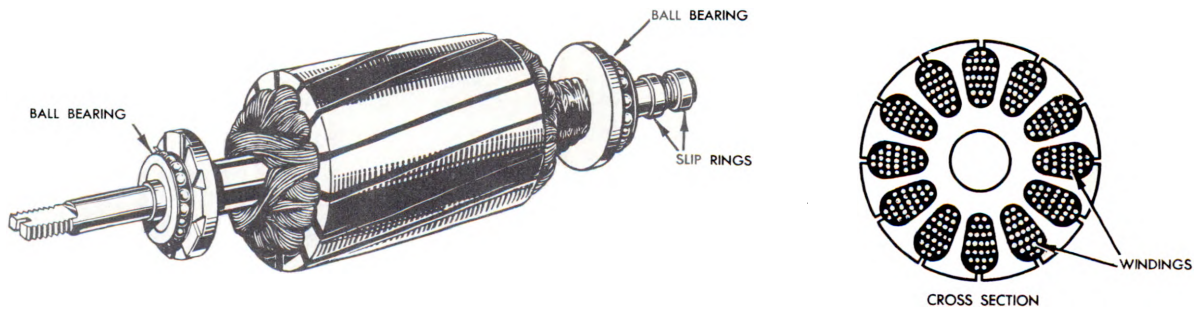


Figure 14-37. Typical Control Transformer Rotor

### The Control Transformer

Synchros are used to cause two shafts to rotate in synchronization and to produce an error voltage that indicates the difference in position of two shafts. The systems discussed up to this point are for synchronization purposes. For the purpose of producing error voltages, there is a type of synchro known as the control transformer that is used in conjunction with a synchro generator. Both the generator and the transformer have their shafts connected to loads. The voltages induced in the stator coils of the generator are applied to the stator of the control transformer. These voltages induce in the rotor of the control transformer (not connected to the 115v AC) a voltage whose magnitude and polarity depend upon the relative positions of the two shafts.

The stator of a control transformer is very similar to that of a synchro generator or motor except that the coils are wound with more turns and finer wire. The purpose of this is to make the impedance high enough to prevent high currents from flowing. Remember that no voltage is induced in the stator coils of the control transformer to buck the applied voltage and keep the current within safe limits. The stator current is determined by the impedance of the stator windings and is practically independent of the rotor's position.

The rotor of a control transformer (Figure 14-37) looks somewhat like the rotor of a differential synchro. Yet there are some important differences. Instead of three groups of windings spaced  $120^\circ$  apart, all the windings are in series and have their external connections through two slip rings and brushes. The shaft always connects to a load making an oscillation damper unnecessary. The coils contain many turns of fine wire. The turns ratio is such that the maximum

output voltage, with normal stator voltages, is 55 volts. There is no appreciable current in the rotor windings. The rotor is not free to turn to any particular position when voltages are applied to the stator windings. The electrical zero position of the rotor is such that no voltage is induced in the rotor windings. This means that the coil axis is at the right angle position as shown in Figure 14-38. Note that the terminals are labeled exactly as are those of a synchro motor or generator.

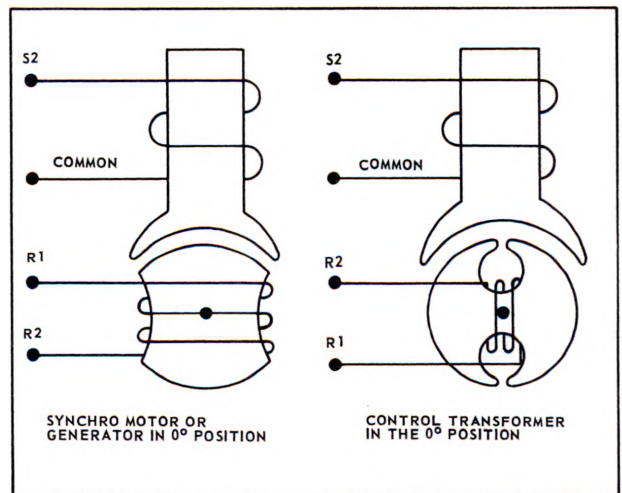


Figure 14-38. Comparison of Electrical Zero Position

The control transformer operates in this manner. In Figure 14-39, the generator shaft is at  $0^\circ$  and the control transformer shaft is at  $0^\circ$  position. The axis of the magnetic field set up by the stator voltages is in the  $0^\circ$  direction — at right angles to the rotor windings. It induces no voltage in the rotor windings. Now assume that

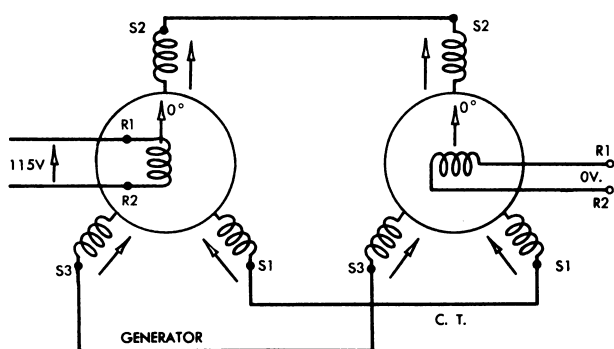


Figure 14-39. Rotor at 0° Position

the control transformer rotor is turned to the 90° position. The axis of the coil will then be aligned with the composite magnetic field and there will be a maximum (55v) voltage induced in it. The phase polarity of the R1-R2 voltage will be identical to the R1-R2 voltage of the generator rotor.

At other positions you can determine the magnitude of the induced voltage by resolving the magnetic field into two components, one of which is perpendicular to the axis of the coil and induces no voltage, and the other which is aligned with the coil and induces a voltage. The component of the field that is parallel to the coil axis is proportional to the sine of the angle to which the control transformer rotor is turned. Note in the induced voltage vs. shaft position graph, Figure 14-40, that voltages above the zero axis are in phase with the generator rotor voltage,

R1-R2, and that voltages below the axis are 180° out of phase. Obviously, then, the voltages from the 180° position to the 360° position are 180° out of phase.

When the rotor of the generator is turned to some position other than 0°, the magnetic field of the control transformer will follow and assume the same position. When the rotor of the control transformer is turned to the same position, the output will be zero volts, since the axis of the coil is perpendicular to the field. The rotor, starting from this position, on rotating in a counterclockwise direction will cause the voltages induced in the control transformer to assume the values indicated in Figure 14-40, where the abscissa is the difference in angular position of the two rotors. This action indicates that if the control transformer (CT) shaft *leads* the generator shaft by any number of degrees up to 180, the output is *in phase* with the generator rotor voltage, but that if the CT shaft *lags* the generator shaft by any number of degrees up to 180, the output is *opposite in phase*.

## SERVOMECHANISMS

You have already seen how synchros transmit data from one place to another. When the shaft to be driven at the remote position is connected to an indicating device or some light load, the synchro motor is capable of developing the necessary torque. If the load is greater and more torque is necessary, then torque amplification is

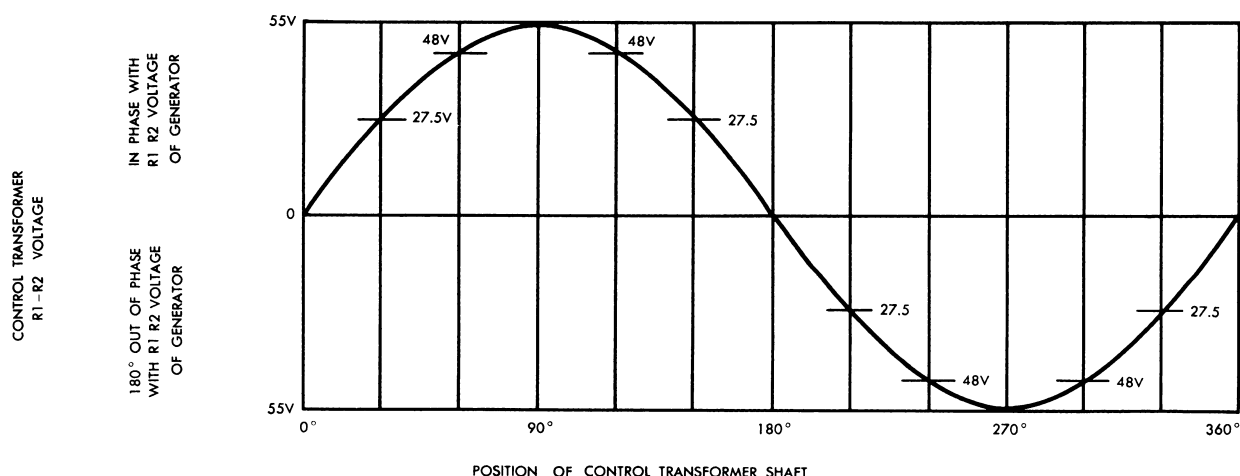


Figure 14-40. Induced Voltage vs. Shaft Position

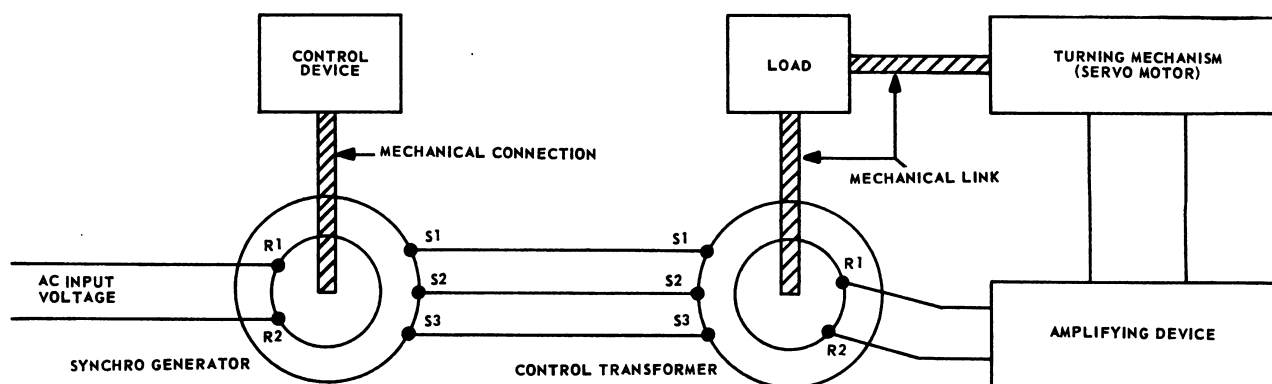


Figure 14-41. Simple Servomechanism

required. In this case the load is driven by a reversible adjustable-speed motor which is called a servo motor. The complete system, including the motor and its control devices, is called a *servomechanism*. The speed and direction of rotation of the motor are controlled by the magnitude and phase of an “error voltage” that is produced by any misalignment of the two shafts (the load shaft and the control shaft). Obviously, the direction of rotation must be such as to reduce the misalignment causing the error voltage.

The principles of the servo system in its simplest form are shown in Figure 14-41. Assume that the control device is a crank or a handwheel and the load is the antenna. The handwheel is connected to a synchro generator, and the antenna to a control transformer. The stator windings of the generator and control transformer are connected electrically as shown. The output of the control transformer is an AC voltage that depends upon the relative shaft positions for magnitude and phase. There is no output when the shafts are in the same position, and accordingly no output from the amplifying device. Hence, the servo motor does not turn and the antenna remains stationary. To change the position of the antenna, you merely turn the handwheel to the desired position. This produces an error voltage in the control transformer. This voltage is rectified and increased in magnitude by the amplifying device and then applied to the servo motor, causing it to turn the antenna to the desired position.

The synchro generator and control transformer may be interchanged in position. However, such an interchange necessitates five electrical leads

from one position to the other and gives no advantages. If the control device rotates continuously in one direction, the servo motor will rotate in the same direction at the same rate but with a small angle of lag. Remember that an error voltage is necessary before torque is developed. The greater the torque required, the greater will be the angle of lag.

### The Servo Motor

An AC motor, inherently a constant speed device, is not suitable for the requirements of a servomechanism and therefore a DC motor is used. There are two methods of controlling the speed and direction of rotation — controlling the voltage applied to the field coils, and controlling the armature voltage. Controlling the field voltage permits control of speed only from a certain minimum speed upward, since speed varies inversely as the flux set up by the field. However, reduction of speed to zero at times is necessary; hence, the control in the servo system is always accomplished by varying the armature voltage.

Placing a variable resistor in series with the armature to control the armature voltage is wasteful since the resistor itself dissipates considerable power. Usually armature voltage is controlled by a DC generator driven by an AC motor. The Ward Leonard drive, discussed later, is an example of this system.

Servo motors operate with constant fields. The field is supplied either by permanent magnets or by DC applied to the field coils. The method of applying DC voltage to the field coils is wasteful of power, since power is consumed even when the motor is not turning. In addition, the design of

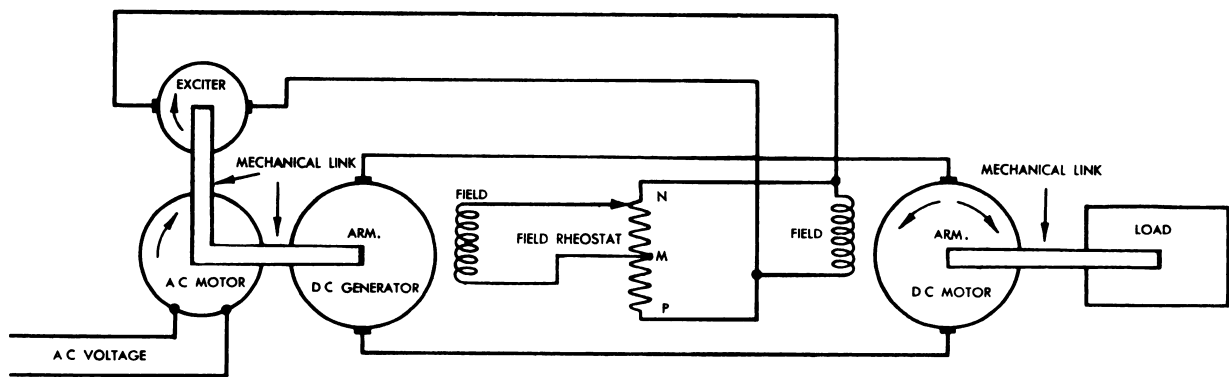


Figure 14-42. Ward Leonard Drive

the equipment is much more complicated than the permanent magnet arrangement. Mainly for this reason the permanent magnet is the more widely used.

One problem in the use of permanent magnets is the danger of demagnetization of the magnet by the field set up by current flow in the armature. This is overcome by the inclusion of a special winding that is connected in series with the armature and wound in the opposite direction. In this winding, called the compensating winding, a compensating field is set up that results from current flow through it. Since the current in the compensating winding varies with the armature current, the compensating field balances out the armature current and protects the permanent magnets.

### The Ward Leonard Drive

In the Ward Leonard drive shown in Figure 14-42, an AC motor turns at a constant speed always in the same direction. It drives the armatures of the large and small DC generators. The small generator provides excitation voltages for both the large DC generator and the motor. The DC motor field is constant, being supplied by field coils. The generator field is not constant. Both its polarity and magnitude are controllable by a field rheostat.

When the variable arm is at the point M, no voltage is applied to the generator field coils and, consequently, there is no output voltage, no voltage applied to the armature of the DC motor, and no motor operation. When the variable arm is between M and N, a voltage of a certain polarity

is applied to the field coils of the generator; the generator output has a certain polarity and causes the motor to turn in a certain direction. When the variable tap is located between M and P, the generator field voltage is reversed in polarity, the generator output is reversed in polarity, and the motor turns in the opposite direction.

In either case the motor speed depends upon the voltage applied to the field coils and is controllable by the rheostat. Thus, you can control both the speed and direction of rotation of the motor. It will therefore function very well as a servomechanism if the generator field is controlled with an error voltage instead of the field rheostat.

### Servo System Using the Ward Leonard Drive

When the Ward Leonard drive is used for a turning mechanism, the servo system operates as shown in Figure 14-43. The error voltage produced by the control transformer (CT) is fed into the control amplifier, whose output is a DC voltage having a polarity and magnitude that is determined by the phase and magnitude of the CT voltage. The output of the control amplifier is applied to the field of the DC generator of the Ward Leonard drive, starting the action just described. The circuits of the control amplifier are explained later. The purpose of synchro generator G2 and synchro motor M1 is to furnish information to the operator as to the position of the antenna. In some cases this could be dispensed with since the scope pattern furnishes the same information when the scan is of the B or the PPI type.



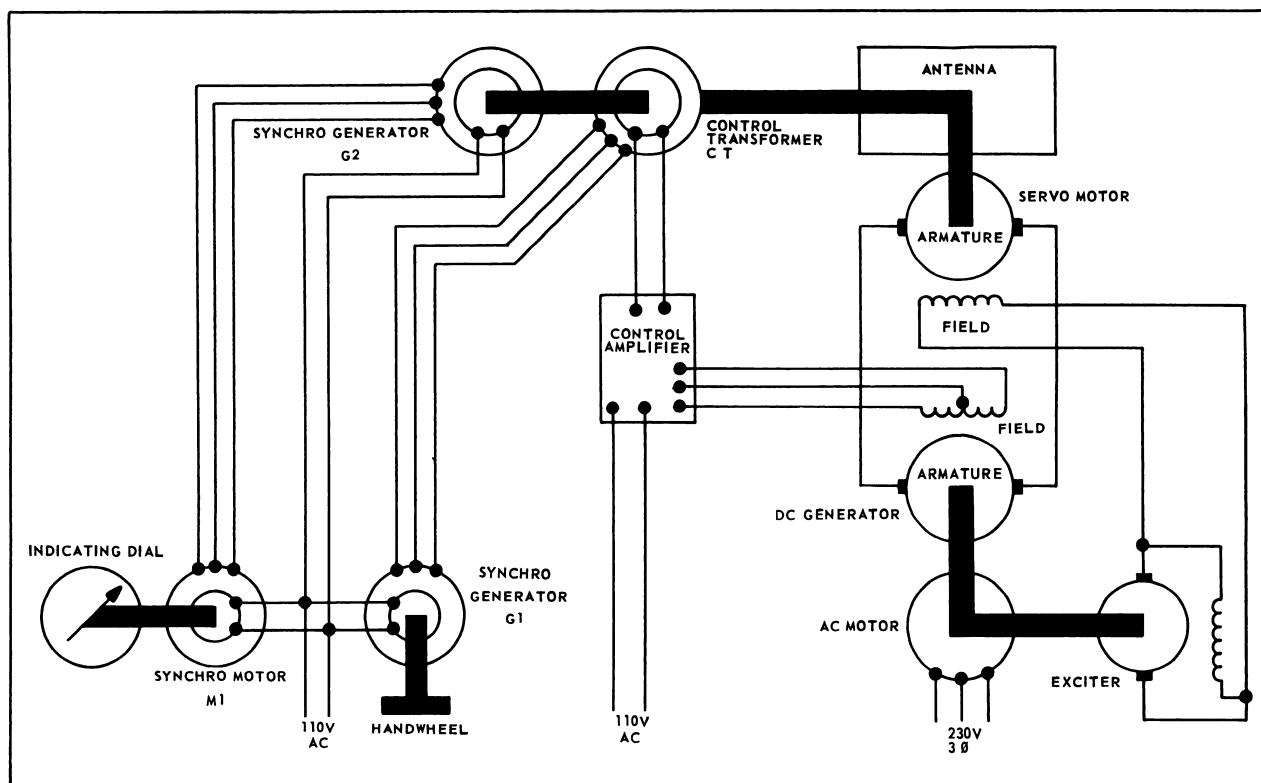


Figure 14-43. Servo System Using Ward Leonard Drive

### The Amplidyne

The amplidyne is a special type of DC generator commonly used for the DC generator in a system similar to the Ward Leonard. Such a system is known as an amplidyne servomechanism. The word amplidyne means dynamo amplifier.

The DC generator may be considered an amplifier since a small amount of power applied to the field coils controls 10 to 100 times as much power in the output of the generator. Of course, the additional power comes from the mechanical power that turns the armature. This mechanical power is comparable to the B+ voltage of a vacuum tube amplifier.

If greater amplification is desired than possible with a single DC generator, then two are used in cascade; that is, the output of the first (the exciter) is the input to the second. While an exciter was used in the Ward Leonard drive system, it may not be considered two stages of amplification, since the input to the exciter was uncontrolled. When two generators are used in

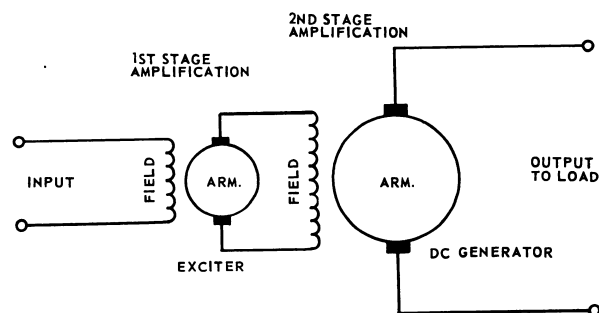


Figure 14-44. DC Generators as Cascade Amplifiers

cascade as shown in Figure 14-44, the over-all amplification is the product of the two amplifications. For example, if each amplifies the power 30 times, the total amplification is 900.

The amplidyne performs as a two-stage DC generator-amplifier with the two stages combined into one armature, Figure 14-45. The input goes to the control field which is on the direct axis of the machine. Rotation of the armature in the control field sets up a voltage at right angles to the field as shown in Figure 14-45.



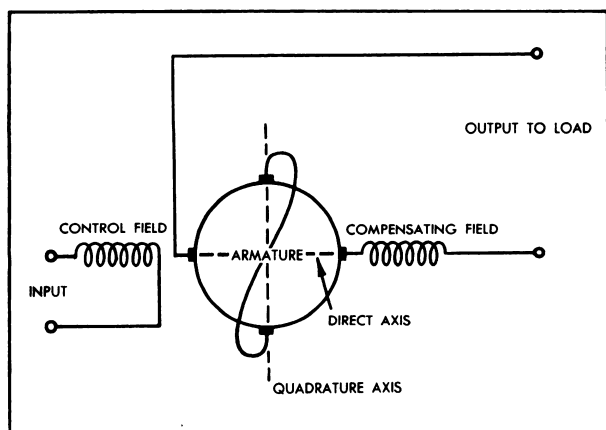


Figure 14-45. The Amplidyne

Brushes are placed on this axis. If a load were connected to these brushes, the machine would become an ordinary DC generator. However, in the amplidyne these brushes are short circuited and quite high currents flow in the armature producing a magnetic field along the quadrature (vertical) axis. This field is much stronger than the control field. This is known as armature reaction, the same as it is in an ordinary generator. The rotation of the armature through this quadrature field sets up a voltage at right angles to it, that is, along the direct (horizontal) axis.

Brushes located on the direct axis have the load connected to them. When the load draws current, a magnetic field is set up along the direct axis because of the armature reaction of this second stage. Since the field due to armature reaction in each stage is  $90^\circ$  from the field producing it, the reaction field of the second stage is  $180^\circ$  from the control field and tends to cancel it out, thereby reducing the amplification of the amplidyne. This is comparable to degeneration in a vacuum tube amplifier. In order to prevent this cancellation of the control field, a compensating winding on the direct axis and in series with the brushes on the direct axis sets up a field that opposes and cancels out the armature reaction of the second stage.

As shown in Figure 14-46A the various magnetic fields of an amplidyne occupy certain positions and have certain directions. The symbol  $\phi_1$  is the control field,  $\phi_2$  the field due to armature reaction of the first stage or the quadrature field,  $\phi_a$  is the armature reaction of the second stage and  $\phi_b$  is the field due to the compensating

windings. Since, by design,  $\phi_b$  equals  $\phi_a$  but is opposite in direction, the net result is as shown in Figure 14-46B.

In order to guard against residual magnetism in the direct axis, a demagnetizing winding called a "killer" winding is sometimes found on the stator. It is excited with alternating current from a magneto generator mounted on the same frame. The magneto consists of a bar magnet revolving within a separate field winding to generate the AC voltage. The alternating current flowing in the stator windings forms a demagnetizing system that neutralizes any residual magnetism when the control field current is zero.

Actually there are two control field windings in an amplidyne when it is used with a control amplifier. The reason is that the vacuum tubes of the control amplifier conduct only in one direction and in order to control the polarity of the output voltage on the amplidyne, it is necessary to reverse the direction of the control field.

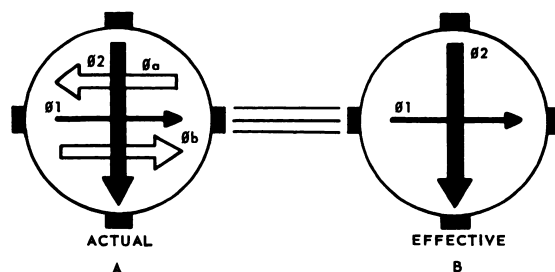


Figure 14-46. Magnetic Fields in Amplidyne

A practically constant input ratio makes it necessary that there be no saturation of the frame and the armature core. The use of a *killer* winding assists in making the gain constant. The power gain of the amplidyne is of the order of several hundreds or thousands and may even be tens of thousands. The full load input is about 0.25 to 0.75 watt, which can easily be supplied by the control amplifier. The output varies from hundreds of watts to several thousand watts. The speed of response to changes in input depends upon the time constants of the control field, the quadrature windings of the armature, and the load circuit or direct axis armature windings. Under the usual operating conditions, the time lag is about 0.1 second.

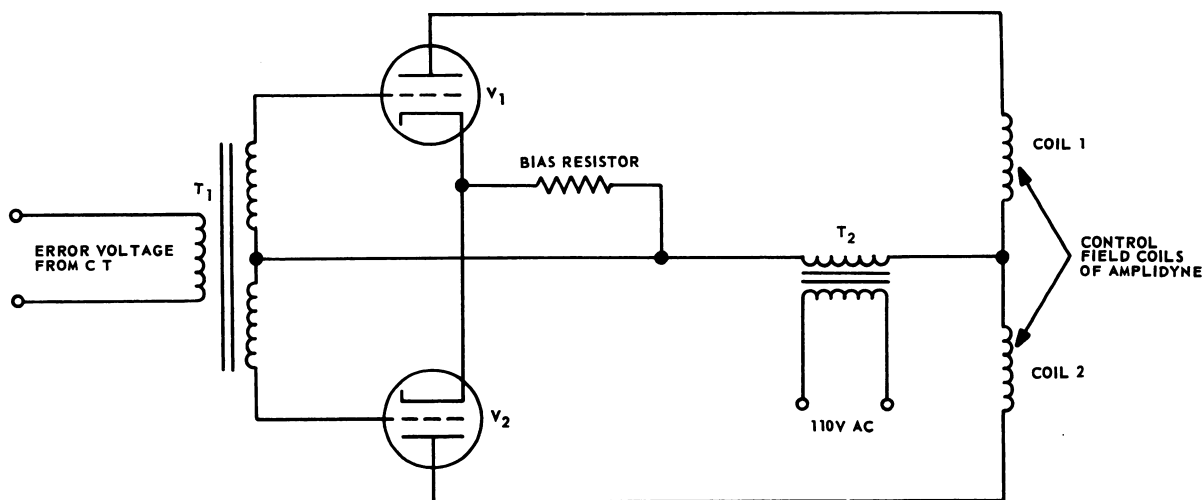


Figure 14-47. Basic Control Amplifier Circuit

### The Control Amplifier

As mentioned previously, the function of the control amplifier is to take the error voltage produced by the control transformer and to convert it into a DC voltage. The magnitude and polarity of this voltage is determined by the magnitude and relative phase of the error voltage. In the control amplifier circuit in Figure 14-47 the two triodes are operated in push-pull, and an AC voltage from a 110-volt line is applied to the plates through transformer  $T_2$ . Bias for the tubes is provided by a common cathode resistor. The error voltage is applied to the grids through transformer  $T_1$  in such a way that the two grids are  $180^\circ$  out of phase. It is either in phase with or  $180^\circ$  out of phase with the 110-volt line voltage depending upon the relative position of the two shafts (synchro generator and control transformer).

To follow through the operation of the control amplifier, assume that the error voltage is in phase with the 110v and that the grid of  $V_1$  goes in a positive direction at the same time that the plate goes positive. This causes an increased current flow through  $V_1$  and this, in turn, an increase through coil 1. At the same time the grid of  $V_2$  has gone in a negative direction, and there is a decrease in current through  $V_2$  and coil 2. Then in the amplidyne the field set up by coil 2 will decrease.

There will then be an output voltage of a certain polarity from the amplidyne causing the

servo motor to turn in a certain direction. During the alternation of the 110v when the plates are negative, no current flows through either tube. Hence the current through coils 1 and 2 is pulsating direct current. (Figure 14-48 shows current and voltage relationships.) If a greater voltage is applied to input transformer  $T_1$ , it will result in more current passing through coil 1 and less through coil 2, which causes the servo motor to turn faster.

When the error voltage is  $180^\circ$  out of phase with the 110 volts, the grid of  $V_2$  will go positive and the grid of  $V_1$  negative at the time the plates go positive. As a result, the current flow through coil 2 will increase and that through coil 1 will decrease. This causes the flux in the amplidyne to increase but in the opposite direction so that the output voltage is opposite in polarity. The servo motor will turn in the opposite direction.

In actual practice some sort of filter circuit is employed to smooth out the pulsations of the control amplifier output. This makes the output steadier. One way of accomplishing this is to connect resistors in parallel with the amplidyne control field windings. During the negative alternation of the plate voltage, the self-induced voltage of the inductance of the control field maintains the field current, which finds a complete circuit through the shunting resistor. Another way of making the output more nearly pure DC is to put an RC circuit in to act as the load impedance and use this practically constant voltage to feed into a DC amplifier.

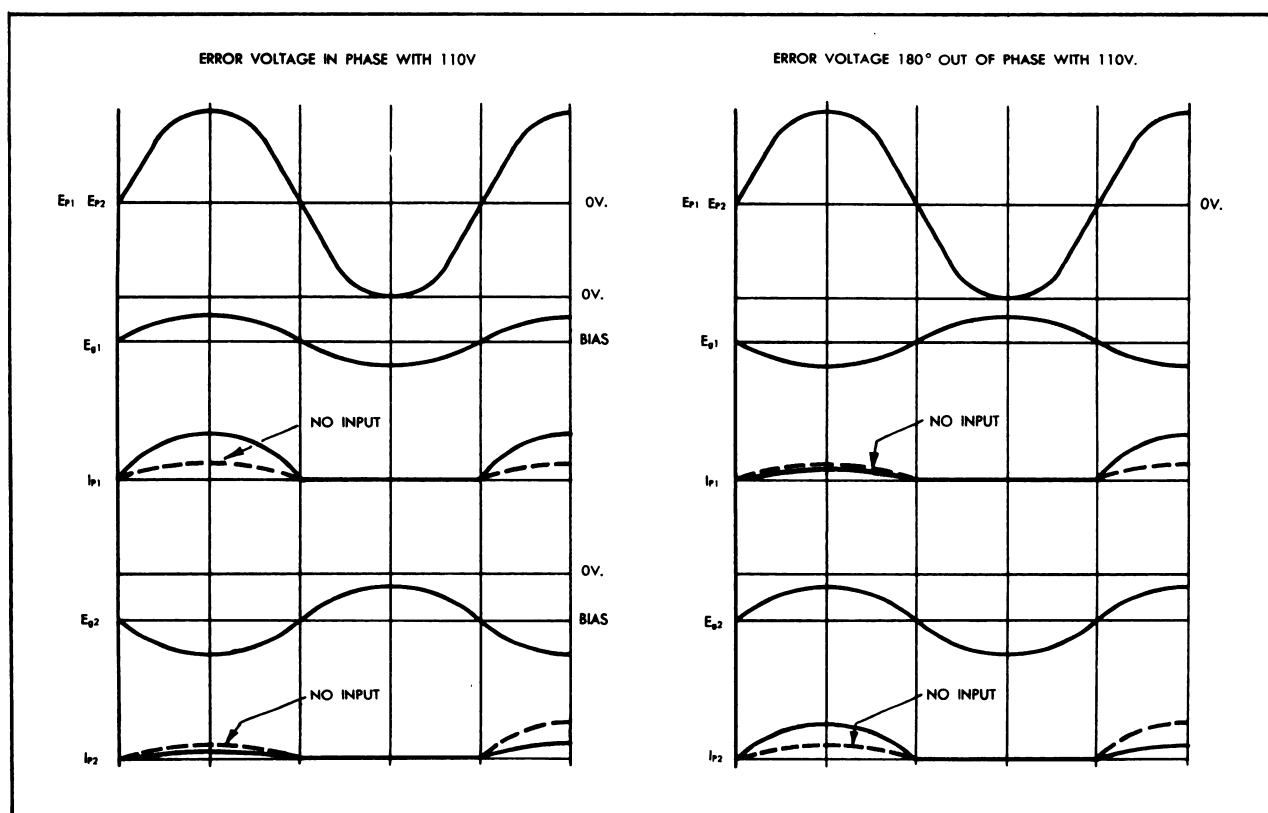


Figure 14-48. Control Amplifier Currents and Voltages

*Hunting* is quite likely to take place in a servomechanism unless some method is devised to prevent it. Hunting means that the servo motor drives the load past the desired position causing an error voltage that reverses the motor's direction; the motor will swing back and forth past the final position several times before stopping. Under some conditions continued oscillation may take place.

The usual anti-hunt circuit consists of feeding back a portion of the amplidyne output into the input of the control amplifier. This voltage is proportional to the speed of the motor and is applied so that it makes the grid of the conducting tube more negative (in other words, it decreases the gain of the amplifier) and decreases the amplifier output and the motor speed. In this way the motor, in effect, acts as a brake on its own speed. The magnitude of the feedback decreases as the speed decreases and allows the motor to rotate the load to the desired position without unnecessary hunting. The amount of feedback is adjustable by a potentiometer.

## USE OF SERVOMECHANISMS FOR AZIMUTH STABILIZATION

The heading of an aircraft is quite changeable. For this reason, it is sometimes desirable for the presentation of data on the indicator scope to remain in the same position, regardless of changes in heading of the aircraft. For example, if the 0° position of the PPI scan is always true north it is much easier to compare the scope picture with the map of the region. In order to accomplish this, it is necessary to change the relative positions of the antenna and sweep on the scope by an amount equal to the change in heading. This can be effected by applying azimuth stabilization to the antenna drive assembly.

### Synchro System Employing Azimuth Stabilization

The circuits shown in Figure 14-49 are those of a typical set employing stabilization. Stabilization is obtained by shifting the sweep coils of the CRT whenever the aircraft changes its heading. The antenna is driven by the azimuth motor

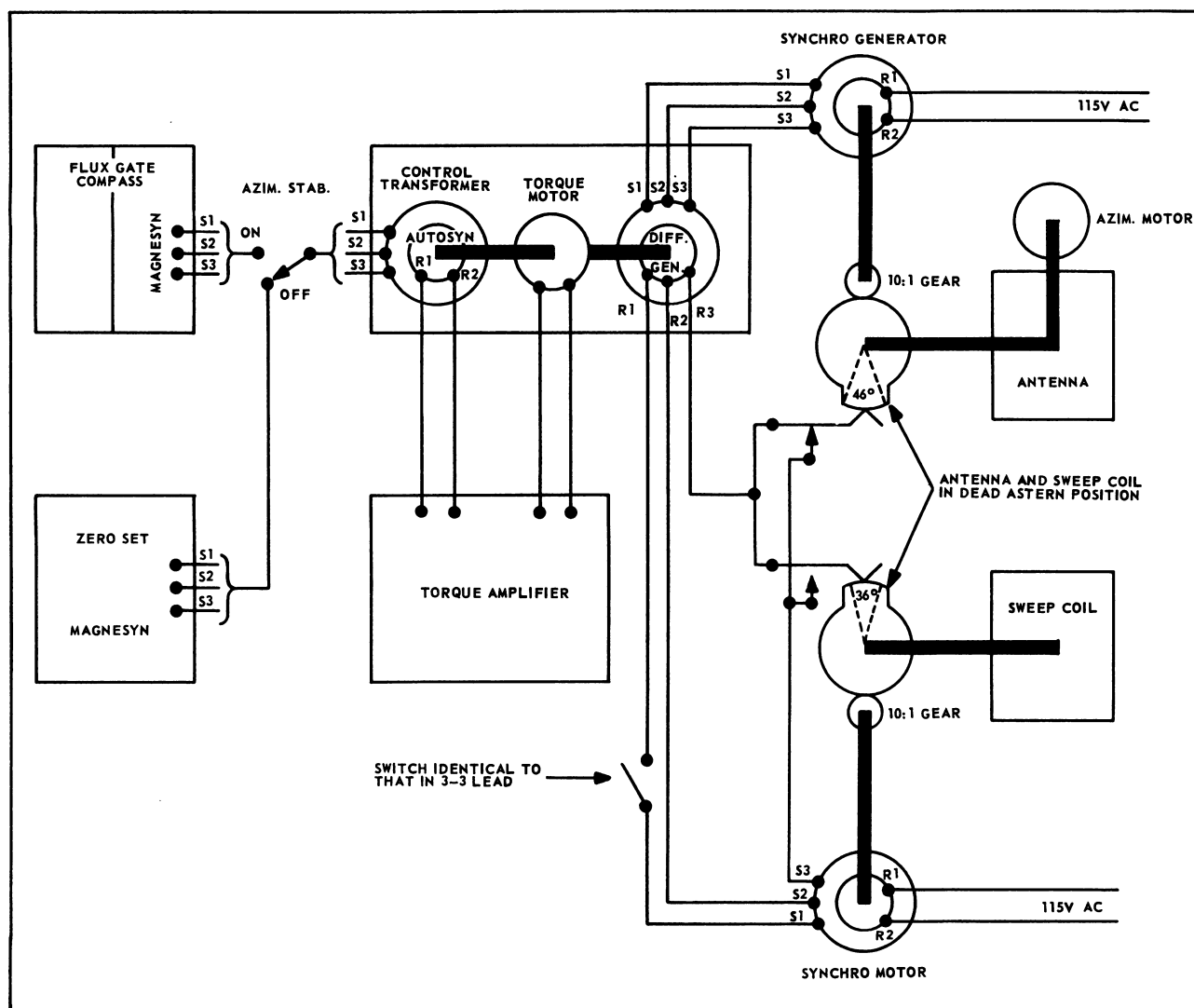


Figure 14-49. Typical Synchro System Using Azimuth Stabilization

at a rate of approximately 24 rpm. A synchro generator is geared to the antenna through a 10:1 gear. The generator is connected to a differential synchro which has its output connected to a synchro motor. This motor is geared, through another 10:1 gear, to the sweep coils and causes them to rotate about the neck of the CRT. If the differential generator is in the electrical zero position, the rotation of the sweep coils is in synchronization with the rotation of the antenna.

The purpose of the 10:1 gears is to reduce to  $\frac{1}{10}$  of its value the lag inherent in a synchro system. In this system, if the synchro motor lags the generator by  $10^\circ$ , it produces only  $1^\circ$  lag between the antenna and sweep coils. The actual

error is a small fraction of  $1^\circ$ . The increased accuracy is gained by the possibility of the antenna and sweep coils being  $36^\circ$ , or some multiple thereof, out of synchronization. Of course the synchros are always in step but since  $36^\circ$  rotation of the antenna (or sweep coils) will mean one complete rotation of the synchro motor, there is the possibility that this error could occur.

To guard against such an occurrence, cams are attached to the shafts of the antenna and the sweep coils. These are called phasing cams and operate switches in the R1-S1 and R3-S3 leads from the differential generator to the synchro motor. The cam attached to the antenna shaft closes the two switches during  $46^\circ$  of rotation

centered about the dead astern position. The cam attached to the sweep coil shaft *opens* switches in the same leads during  $36^\circ$  of rotation, also centered about the astern position. If the two shafts are in synchronization, the antenna cam closes the switches  $5^\circ$  ahead of the position where the sweep coil cam opens the switches and holds them closed for  $5^\circ$  past the position where the sweep coil cam again closes its switches. Thus, the circuits are complete at all times and the synchro motor drives the sweep coils in step with the antenna.

### Operation

To illustrate the action, assume that when the system is turned on and the antenna is pointing somewhere in the aft sector the sweep coils are at a position corresponding to dead ahead. Both sets of phasing switches are closed, and the antenna and sweep coils start rotating. There is large error between the antenna and sweep coils. Figure 14-50 shows the cams and switches at successive steps in the rotation. This out-of-step rotation will continue until both switches are open at the same time (see C), then the sweep coils will stop and remain in that position ( $180^\circ$ ) until the antenna reaches the position where it will close the switch. This occurs at  $23^\circ$  before the dead astern position. The sweep coil rotor

will then electrically interlock and cause the antenna and sweep coils to be in step once more.

The description of the operation thus far assumes that the differential generator was in the electrical zero position. This is true if the AZIM. STAB switch is in the OFF position. In this type operation, the control transformer in the phasing unit has its stator windings connected to a zero set magnesyn. The magnesyn is a device similar in performance to a synchro generator. The control transformer in this particular setup is known as an autosyn. The output of the control transformer, after being amplified by the torque amplifier, causes the torque motor to turn the CT rotor and differential synchro rotor to the desired position which is the electrical zero position of the differential generator.

When using azimuth stabilization, turn the switch AZIM. STAB to the ON position. This disconnects the control transformer from the zero set magnesyn and connects it to the flux gate compass magnesyn. The switch also shorts out the phasing switches and connects the differential generator directly to the synchro motor. The magnesyn in the flux gate compass produces voltages to set up a field in the control transformer, which is aligned with the position of the rotor in the magnesyn. This position is dependent upon the position of the compass. The error voltage produced by the control transformer, after

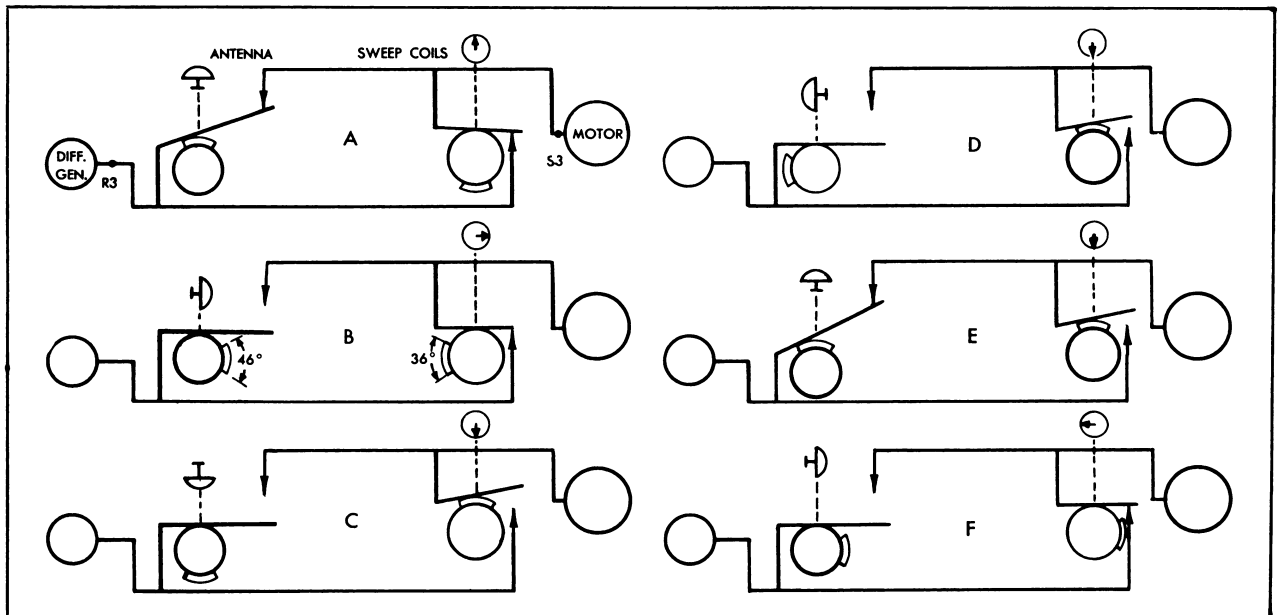


Figure 14-50. Phasing Switch Action

amplification, turns the control transformer rotor and that of the differential synchro to the position corresponding to the compass position.

Since the rotor of the differential generator is shifted by an angle equal to the difference between the aircraft's heading and true north, its output voltages will be such as to cause the sweep coil motor to lead or lag the antenna generator by the same angle. By proper connection and adjustment the sweep coils move in such a way that the 0° position of the scope is true north. In order to determine the aircraft's heading from the scope pattern a switch connected to the antenna causes the bias of the CRT to decrease when the antenna points dead ahead. This results in a brighter trace which indicates the aircraft's heading. The purpose in shorting out the phasing switches is to permit the antenna and sweep coils to be out of step, as is the case when the aircraft's heading is other than north.

### MAGNETIC AMPLIFIERS

Magnetic amplifiers, for the most part, are made up of controlled variable inductances (*saturable reactors*), dry-disk rectifiers, fixed and adjustable resistors, capacitors, and conventional-type transformers. The magnetic amplifier uses the saturation of a magnetic core, which is regulated by the control circuit, to determine the AC circuit resistance. In this way electrical devices can be controlled and electrical signals can be amplified.

Magnetic amplifiers have been in use since the early 1900's. In 1901 a saturable reactor was used in a "wireless signaling" system to amplify the output of a carbon microphone and to modulate a transmitter. During this same period, a saturable transformer was being used to measure currents from 100 to 1000 amperes with an accuracy of about  $\pm 0.5\%$ . In 1915 the development of electronic amplifiers did away with the magnetic amplifier.

Later, experimentation with magnetic amplifiers introduced feedback, push-pull circuits, and self-saturating circuits. From 1930 on, much of the work was for the military. Military application required unusual durability, simplicity, and reliability. New core materials and efficient and reliable dry-disk rectifiers brought about still further development in magnetic amplifiers.

Reliability is the main advantage of the magnetic amplifier. There are no moving parts, no contacts, no filaments or other features that are prone to fail in electron-tube amplifiers. Magnetic amplifiers are sturdier than electron-tube amplifiers; they absorb vibrations and shocks easily. Since they have no filaments, they require no warm-up time and operate the minute the switch is turned on. Magnetic amplifiers are sturdy; since they have no moving parts, they do not require servicing or replacement. They can be hermetically sealed. Magnetic amplifiers are suitable for measuring or amplifying extremely small voltages or currents.

The size of the core of the saturable-reactor elements in a magnetic amplifier is inversely proportional to the power-supply frequency. When a 400-cycle supply is used, the weight and size of the magnetic amplifier can be less than that of a vacuum-tube amplifier for the same circuit. Since the output of the magnetic amplifier can be completely isolated from the input, multiple input windings can be used. This allows the simultaneous application of many control signals without electric interference among them. Power gains of 10,000 to 200,000 in a single-stage circuit can be obtained with magnetic amplifiers. The range of input impedances that can be matched by the control winding is from 0.1 ohm up to about 100,000 ohms. The output impedances that can be matched by the load circuit range from about 5 to 5000 ohms with a power supply frequency of 60 cycles per second.

Disadvantages of the magnetic amplifier are poor frequency response and the long time of response.

Magnetic amplifiers, vacuum-tube amplifiers, and transistor amplifiers should not be considered as rival devices. They may be combined to take the greatest advantage of the properties of each. Certain types of amplifiers are better than others in each application; the one best suited for the job should be used.

### SATURABLE REACTORS

In order to understand the operation of a magnetic amplifier, it is necessary to understand each of the components in the circuit. All except the saturable reactor have been discussed in previous chapters.



A saturable reactor, or *transductor*, consists of a closed magnetic core on which are mounted one or more load windings and one or more control windings. Since there are no moving parts, it is one of the simplest of electrical devices. Properly designed and constructed, the saturable reactor requires practically no maintenance and has an extremely long life expectancy.

The core material for saturable reactors requires certain magnetic properties. Losses due to hysteresis and eddy currents must be minimum. Resistivity must be high, and coercive force low. Saturation flux density should be high so that a given weight of core material will have large power-handling capabilities. The hysteresis loop should be nearly rectangular. The core material must also have stable magnetic characteristics under conditions of shock, of changing temperature, and of changing magnetic strain. It must also be possible to form the material into thin tapes or laminations. A lamination is a thin sheet of the metal from which the core is made.

### Construction

There are many types of core constructions used in saturable reactors. A ring core (toroid) may be made of a stack of flat circular-shaped laminations with a circular opening in the center. The material is preferably nondirectional in magnetic characteristic. The most common core used in the better saturable reactors is made by winding a tight spiral from a thin strip of grain-oriented core material. For best results the direction of orientation should be parallel to the long side of the strip. (Grain-oriented materials magnetize more easily in one direction than in others.)

Insulation can either be wound on the core or be in the form of cylinders to fit inside and outside the core, with a ring-shaped collar at each end. The latter type of insulation is used where the voltage between winding and ground is below 1300 volts, because of the danger of arc-over at the joint between the cylinders and the collars.

Rectangular cores are often made of rectangular laminations that can be stacked into the pre-wound coil. Two-legged cores are made from U- and I-shaped laminations, the I-shaped pieces fitting across the end of the U-shaped piece. Three-legged cores made from E- and I-shaped pieces of core material are put together with the I-shaped piece at opposite ends of the core in adjacent laminations. For best performance, the

air gap must be as small as possible and the reluctance kept to a minimum. The usual interleaving technique used in ordinary transformers is unsatisfactory, as the many small air gaps provided by the butt joints of the standard E-core distort the magnetization curve.

An improvement is the four-legged core with the openings between the legs in the laminations punched out and only the two inner legs loose at one end. The AC windings are slipped on the two inner legs with the control windings around both. When the windings are on the legs, the legs are bent back into position. The very small joint in this type of core gives it an advantage over the other rectangular cores. However, the joints are not normally perfect because of the oxide coating or other insulation on the edges of the pieces at the joint. This adds to the reluctance of the core.

Three-phase saturable reactors can be made from three toroidal (ring) cores, each with one winding, by stacking them and wrapping the control winding around all three. Another type of three-phase saturable reactor consists of a three-legged core with a load, or gate, winding on each leg and a control winding around all three legs. Various other types of construction may also be used.

The physical arrangement of the various toroidal windings on the cores will greatly affect the performance of the magnetic amplifier. They must be wound in a way that will reduce leakage effects to a minimum and minimize the circulation of alternating current in the DC control, bias, and feedback windings.

Regardless of the size of the saturable reactor, it controls some device or process that has dependent on it huge sums of money, or even human life. For this reason, protection of the elements of the saturable reactor is very important. The insulation, protection, and enclosure must be evaluated, not on the value of the reactor itself, but on the value of the process it controls.

### Classification

The saturable reactor permits control of large amounts of electric power with much smaller control power. This means that smaller control equipment can be used than when the load is controlled directly. The saturable reactor and the load are placed where they are needed; the

control equipment is placed at a convenient location without excessive loss in transmission between the control and the load.

Magnetically controlled saturable reactors can be divided into two main groups — those controlled by a constant magnetomotive force, and those controlled by a variable magnetomotive force. Under the first classification come frequency transformers and constant current devices, which use either a constant magnetic field or a constant electric current. The second group contains devices using a variable magnetic field, such as various current and voltage measuring devices and magnetic amplifiers. (The main difference between magnetic amplifiers and flux-gate magnetometers is that in the amplifier the saturable reactor is controlled by the electric current or voltage to be amplified, whereas in the flux-gate magnetometer, the saturable reactor is controlled by the magnetic field to be measured.)

### Principles of Operation

During steady-state operation the saturable reactor may be looked at as an impedance to the AC circuit that is very high through one part of the half-cycle of the supply voltage and very low during the rest of the half-cycle.

As you know, the inductance value of a particular coil depends upon its physical characteristics. An air-core coil can have its inductance increased by the insertion of a magnetic core, since inductance is directly proportional to the permeability of the core. This means that if an iron-alloy core with a permeability of 1250 is put in an air-core coil, the inductance becomes 1250 times greater than the value with the air core.

As the iron core is moved in and out of the winding, it alternately increases and decreases the effective permeability of the coil's core. Changing the permeability changes the inductance. This will change the inductive reactance of the coil. Figure 14-51 shows an air-core inductance in series with a load and a source of power. The inductive reactance is minimum and the current is maximum. The greater part of the applied voltage is dropped across the load. When the iron core is moved into the coil, the inductance, and thus the inductive reactance, increases. As a result, the current decreases, and the voltage drop across the load decreases proportionately. Moving the iron core in and out of the coil will alternately increase and decrease

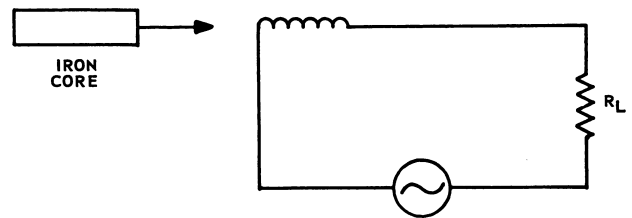


Figure 14-51. Inductance

the inductance and, as a result, decrease and increase the voltage across the load.

A comparatively small amount of energy is enough to control the output of the source, which may have a rating of several horsepower. It is for this reason that the circuit acts as a magnetic amplifier.

A more practical way of varying the permeability of the core is shown in Figure 14-52. The winding in series with the load has a fixed magnetic core. The core's permeability is varied by saturating or unsaturating the material with a relatively small current applied to the control winding.

To better understand how the circuit works, let us first review some of the basic magnetic concepts. When current flows in a coil, it produces a magnetomotive force, which can be compared to the electromotive force in an electric circuit. The strength of the magnetomotive force depends on the ampere turns (current flow times the number of turns in the coil).

The mmf produces magnetic flux, which consists of closed lines of force, that is comparable to current in an electric circuit. Like current, the flux varies directly with the force but indirectly with the opposition (reluctance) of the path making up the magnetic circuit. Magnetic reluctance is a property of the material of the

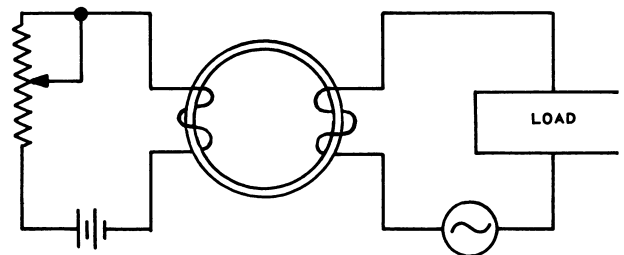


Figure 14-52. Load Control

magnetic path. In an electromagnet the value of the reluctance depends primarily on the core.

The reluctance of an air core is constant, but the reluctance of a ferromagnetic material can change with the change in flux. In an iron-core coil, when the magnetizing current first begins to flow, the reluctance is low and the flux is very high compared with that in an air-core coil under similar conditions. The reluctance will increase if the current is increased, but the rate-of-flux increase falls off.

As the current increases further, the reluctance increases still more rapidly, its value approaching the reluctance of air. When further increases in the magnetizing current produce very little increase in the total flux, the core is said to be saturated.

Figure 14-53 is the permeability curve for the ferromagnetic material used as core in Figure 14-52. From this graph you can see that permeability ( $B/H$ ) is not constant. It has different values for different magnetizing forces. Between points 1 and 2 the curve is nearly a straight line. The permeability is high and the coil has a large value of inductance. Between points 2 and 4, a change in the magnetizing force ( $H$ ) produces a much smaller change in the flux density ( $B$ ) than the same change in  $H$  for the linear portion of the curve. The permeability between points 2 and 4 is much lower; therefore, the inductance is also smaller. Beyond point 4 the curve begins to flatten out. As  $H$  is increased,  $B$  increases very little, if any. The core is saturated and the inductance of the coil is minimum.

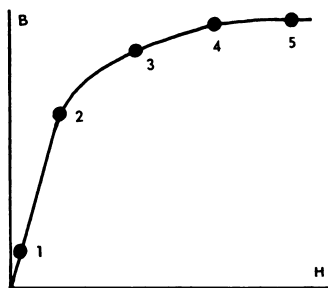


Figure 14-53. Permeability Curve

The control action of a magnetic amplifier depends on change in inductance. If the control signals vary the magnetization of the core above and below point 3 (from 2 to 4), small changes

in control voltage (input) result in comparatively large variations in the output circuit impedance and the output voltage. The control current in Figure 14-52 is adjusted by a potentiometer. This current sets up a flux in one direction through the coil with an intensity determined by the ampere-turns of the control circuit. The load winding is connected in series with the load resistance and an AC power source.

An increase in control current lowers the inductance of the load winding because the magnetization of the core is shifted toward point 4 on the curve in Figure 14-53 where the permeability is less. This lowers the inductive reactance of the load winding. The total impedance of the load circuit decreases; the load current increases, and the power in the load is increased. If the increase in control current is enough to saturate the core, the reactance of the load winding drops to an extremely low value, and maximum supply voltage is applied to the load.

Decreasing the control current increases the reactance of the load winding. In this case the magnetization of the core shifts in the opposite direction, toward point 2 on the curve of Figure 14-53, and the permeability increases. When the load coil has maximum inductance, minimum load current flows and the load power is minimum.

This arrangement is very inefficient because of the transformer action between the load winding and the control winding. The induced AC voltage in the control winding alternately adds to and subtracts from the control voltage.

To eliminate the transformer action between the load and the control windings, a three-legged magnetic core may be used. Figure 14-54 shows a device used for controlling large alternating currents. There is an AC winding on each outer leg. These are arranged so that their alternating flux components cancel (dotted line) in the center leg on which the DC control winding is placed. There will be complete cancellation if the AC windings have the same number of ampere turns and are wound in the right direction.

The alternating flux effectively does not pass through the center leg. However, the two flux components are in the same direction in the outer legs. This composite flux at one instant is indicated by the broken line in Figure 14-54. The control winding produces a magnetic flux that magnetizes the entire core. The direction of the control flux is indicated by the solid lines. The

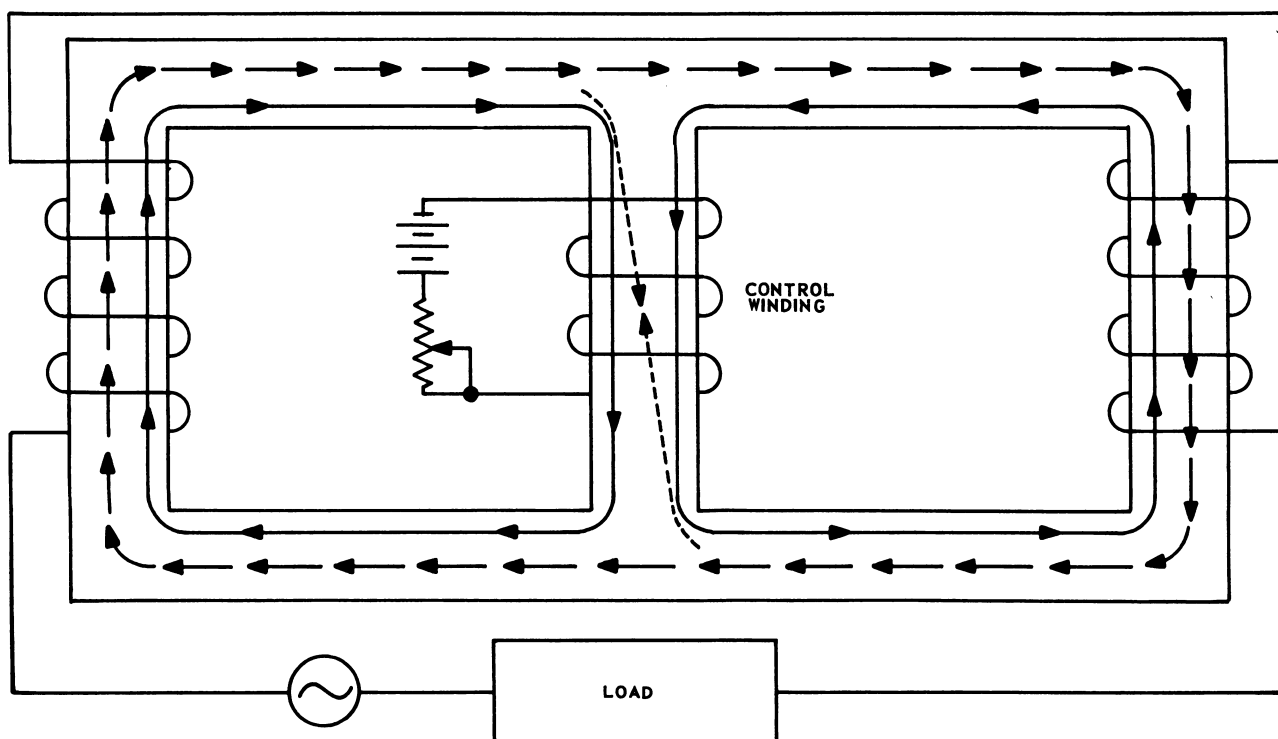


Figure 14-54. Three-Legged Magnetic Core Circuit

DC flux affects the operation of the load circuit, but there is no transformer action from the load circuit to the control winding.

The preferred method of saturable reactor construction uses two separate circular cores that are stacked (Figure 14-55B). Each core has an AC winding. The cores are stacked so that these windings are wound in opposite directions. The control winding is wound around both cores after stacking. Figure 14-55A shows the schematic

diagram of this arrangement. The control winding is shown as two windings connected series aiding.

At any instant the flux in the two cores, due to the alternating current flowing in the load windings, is in such a direction that the induced currents in the control windings cancel each other.

The direction of the flux in the cores at the instant the AC source has the polarity indicated

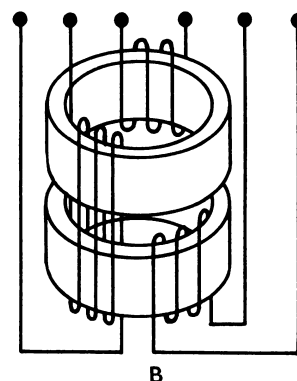
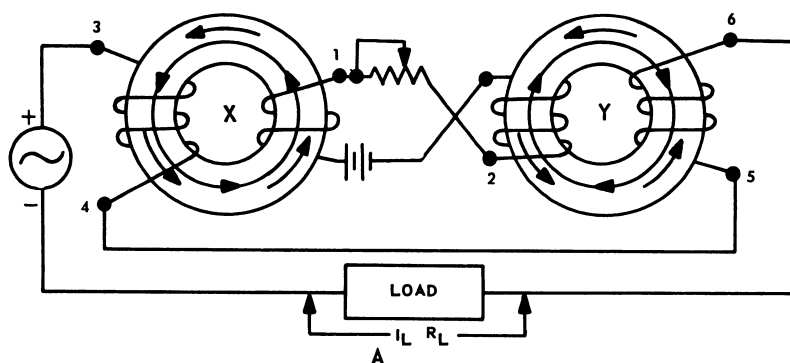


Figure 14-55. Saturable Reactor Control

in Figure 14-55 is indicated by the broken lines. The direction of the flux due to the current in the control winding is shown by solid lines. When the polarity of the AC source reverses, the direction of that flux reverses, but the flux due to the control current continues to be in the original direction, since the control source is DC.

When the control current is minimum, the reactance of the load windings is very high, and minimum load current flows. If the control voltage source is adjusted so that current flows in the control windings, the flux produced aids the AC flux in core X and opposes that in core Y on the half-cycle shown in Figure 14-55A. On the other alternation, the DC flux opposes the AC flux in core X and aids that in core Y. Alternately, core X and core Y are saturated, letting pulses of current flow in the load circuit. When core X is saturated, core Y will act as a transformer. The load winding will induce a pulse of current into the control winding. When core Y is saturated, core X acts as a transformer, and its load winding induces a pulse into the control winding. This pulse of current will be in the same direction as that induced by the winding on core Y.

While building to saturation in each core, the reactance in the circuit is high, and there is little current flow in the load circuit. At saturation there will be no voltage across the windings on the saturated core since the reactance will be zero for all practical purposes. Effectively, this part of the control winding shorts that part on the unsaturated core. When the secondary of a transformer is shorted, the primary appears shorted. Thus, there is no voltage appearing across either core, and almost the entire supply voltage appears across the load resistance.

Figure 14-56 shows the waveshapes of the circuit in Figure 14-55. The upper part of the illustration shows the applied voltage and the voltage drop across the load, due to the load current. The lower part of the illustration shows the hysteresis loop for one core. The type of material in the core produces a very narrow, rectangular, hysteresis loop.

Figure 14-56A shows the waveshape from Figure 14-55 with minimum control current. The voltage drop across the load ( $I_L R_L$ ) is extremely small. In Figure 14-56B the control current is larger, and the core becomes saturated for a part of each cycle. The load current begins to

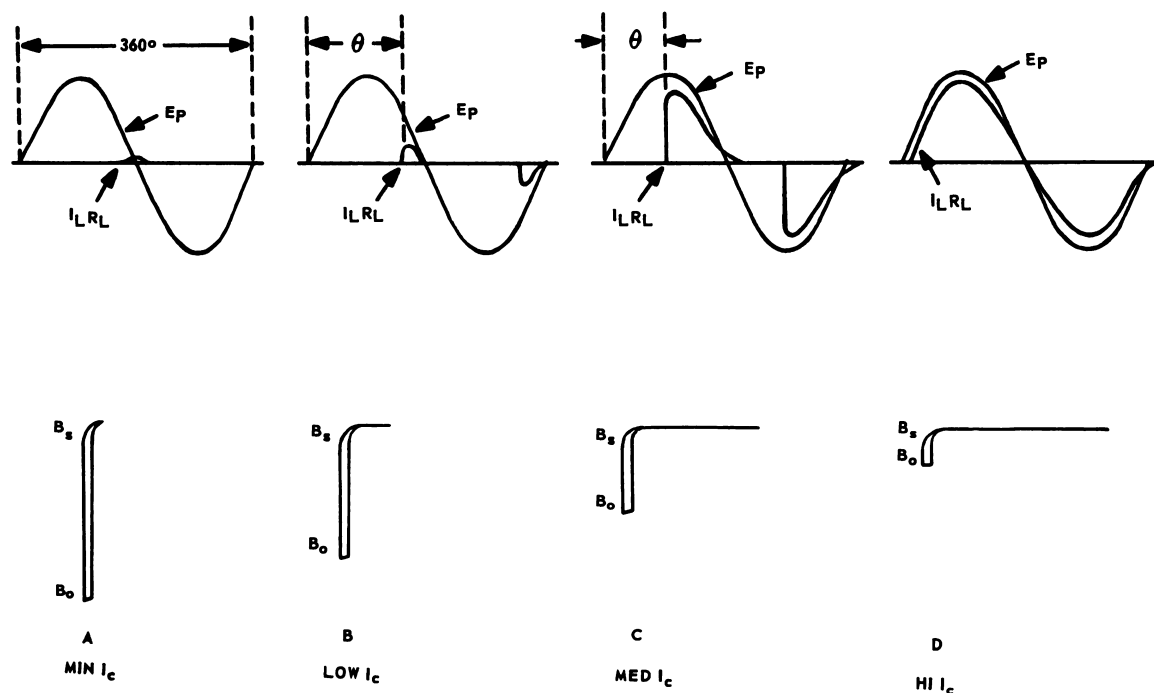


Figure 14-56. Output Voltage and Hysteresis Loops at Various Control Currents

flow earlier and reaches a higher value. The distance along the time axis from the starting point of the applied voltage to the time when the load current begins to flow is the *phase angle*, or *firing angle*. The size of the phase angle depends upon the initial magnetization of the core.

With a steady current flowing in the control winding, the hysteresis loop is much shorter than with no DC at all.  $B_0$ , the minimum flux density (the flux density at the start of each cycle in this arrangement), depends upon the magnetization due to the control current; the higher the control current, the closer to saturation the starting flux density.

In Figure 14-56C the control current is increased to a still higher value than in Figure 14-56B. Notice how abruptly the load current begins. It rises quickly to a maximum, and the voltage across the load at that time is almost equal to the applied voltage. Before the load current starts, all of the applied voltage is across the AC windings; after load current begins to flow, only a very small portion of the applied voltage (because of the resistance of the windings themselves) appears across the AC windings. Figure 14-56D shows the load voltage when the control current is high enough to keep the cores saturated all of the time. The load voltage wave-shape follows that of the applied voltage.

The output voltage waveshapes of Figure 14-56B and Figure 14-56C resemble those of a thyatron with a resistive load. The operation of a saturable reactor can also be compared to that of a synchronous switch. The saturable reactor acts as an open switch between the load and the supply voltage during the first part of each half-cycle when no load current is flowing. During the other half-cycles, the saturable reactor is like a closed switch, allowing current to

flow through the load. The point at which the saturable reactor allows current to flow in the load circuit determines the amplitude of the load, or output voltage. By increasing the control current, you can decrease the phase angle, and thus control the point at which the saturable reactor saturates and also thereby control the amplitude of the output.

One of the disadvantages of the simple magnetic amplifier circuit shown in Figure 14-55 is the requirement for relatively large control currents. The large control current is needed to overcome the effects of the varying magnetic flux caused by the large AC load current. Without any flux due to a direct current flowing in the control winding, the magnetic flux density of the core varies above and below zero in accordance with the alternating current applied by the power source. Thus, during one half-cycle the flux in one core due to the load current will aid that due to the control current, and during the other half-cycle the load current will oppose that due to the control current. This can be considered as feedback. Since the feedback takes place because of the circuit itself, it is called internal, or *intrinsic*, feedback. When the feedback is aiding the input, it is called positive; when opposing the input, it is called negative.

Enough flux must be provided to balance out the oscillating flux of the load current and to set the operating point at the desired place on the magnetization curve. This can be accomplished in two ways. Either the control circuit ampere turns must exceed those of the load circuit, which results in an inefficient amplifier, or else some feedback from an external source must be used to cancel the effect of the changes induced by the load current flux.





## Special Circuits

### DOPPLER SYSTEM

#### Doppler Effect

The Doppler effect, by definition, is the apparent change in the frequency of light, sound, or electromagnetic radiation that is observed when the source and the observer are in relative motion. This frequency shift is external to the transmitter. You have all heard a train's whistle appear to change its pitch from a high tone as the train approaches to a lower tone as it moves away from you. As the train approaches, there is an apparent increase in frequency, a raising of the pitch; as the train moves away from you, there is an apparent decrease in frequency, a lowering of the pitch. This is the Doppler effect.

#### Analysis of the Doppler Principle

What causes this apparent change in pitch? Let us assume that a transmitter emitting an audio signal at a frequency of 60 cycles per second is traveling at a velocity of 360 feet per second. Therefore, at the end of one second the transmitter will have moved from T to T1 in Figure 15-1A. The total distance from T to O, the position of the observer, is 1080 feet. We know that the velocity of sound is 1080 feet per second. Thus, a sound emitted at T will reach O in one second. To find the wavelength ( $\lambda$ ) of the transmitted signal, divide the velocity of the signal (1080 feet per second) by the frequency of the signal (60 cps). The result is 18 feet.

We know that in one second the transmitter moves 360 feet toward the observer and transmits 60 cycles. At the end of one second the

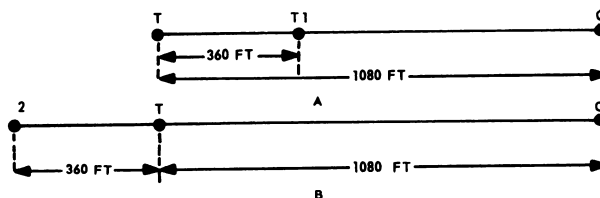


Figure 15-1. Transmitter Moving Relative to an Observer

first cycle of the transmitted signal reaches the observer just as the sixtieth cycle is leaving the transmitter at T1. Under these conditions the 60 cycles emitted are between the observer and T1. Notice that this distance is only 720 feet ( $1080 - 360$ ). The 60 cycles spread over the distance from T1 to O have a new wavelength, which is only 12 feet ( $720/60$ ).

To find the new frequency, use the formula for frequency,

$$\begin{aligned} f &= \frac{V}{\lambda} \\ &= \frac{1080}{12} \\ &= 90 \text{ cps} \end{aligned}$$

The original frequency of 60 cps has changed to an apparent frequency of 90 cps. This new frequency only applies to the observer at O. Notice that the Doppler frequency variation is directly proportional to the velocity of the approaching transmitter. The faster the transmitter moves toward the observer, the greater the number of

waves that will be crowded into the space between the transmitter and the observer.

If the transmitter were stationary and the observer moving, the observer would encounter more waves per unit of time as he approached the transmitter. As a result, he would hear a higher pitch than the transmitter was actually emitting.

When the transmitter is traveling away from the observer, as in Figure 15-1B, the first cycle leaves the transmitter at T and the sixtieth at T2. The first cycle reaches the observer at O when the transmitter reaches T2. There are 60 cycles stretched out over 1080 plus 360 feet, a total of 1440 feet. The wavelength of these 60 cycles is  $1440/60$ , or 24 feet. Their apparent frequency is  $1080/24$ , or 45 cps.

### Continuous-Wave System

Another name which is given to the Doppler system is continuous-wave system. This system, as the name suggests, transmits continuously. The transmitting antenna directs energy to the target, which reflects a small fraction of the energy to the receiving antenna. It is not practical to use a single antenna in the continuous-wave system because the transmitter and the receiver are operating simultaneously and continuously.

These antennas are so oriented that only a small fraction of the transmitted power leaks directly to the receiver. This is done because at microwave frequencies the best practical detector is a crystal, and a crystal would burn out if directly connected even to a low-power transmitter.

### Applications of a Continuous-Wave System

There are several applications of the continuous-wave, or Doppler, system. In one radar application the radar set differentiates between the transmitted and reflected waves, and thus determines the speed of the moving object. The Doppler method is the best means of detecting fast-moving objects that do not require range resolution.

The Doppler effect here can be explained as follows. As a moving object approaches the transmitter it encounters, and therefore reflects, more waves per unit time, just as an observer approaching a transmitter receives more waves per unit time. The amount of the frequency

shift is very small in relation to the carrier frequency, since the velocity of propagation of the signal is very high compared to the speed of the target. However, as the carrier frequencies used in radar are high, appreciable frequency shifts are produced in the audio-frequency range by proportionate speeds of hundreds of miles per hour. One-quarter cycle shift at 10,000 mc will give speed measurements accurate to a fraction of a percent.

The separation of the background and the target is based on this shift in frequency (Doppler effect) that is caused by the reflection of the signal from a moving target. A disadvantage of the system is that it does not determine the range of the target, nor is it able to differentiate between targets when they lie in the same azimuth plane. Moreover, it does not "see" stationary or slowly moving objects, which a pulse system radar can detect.

A second use is to provide information through severe clutter. A pulse system, even with moving target indicator equipment, may have difficulty reading through clutter. Certain Doppler systems can, and do, provide useful and adequate intelligence in such instances.

A third use is involved in air navigation equipment where the continuous longitude and latitude (position) of an aircraft is maintained.

The cost of continuous-wave equipment is less than that of a complex pulse system and, too, this system can work down to zero range, which is something a pulse system cannot do.

### Detection of Doppler Frequency

The detection of the Doppler frequency may be accomplished by various means. The transmitted and received frequencies may be mixed in a detector. The output of this circuit is an audio frequency. In continuous-wave system radars, the output is an audio-frequency sine wave that is heard over a headphone or loudspeaker. In pulsed radars, the output is observed on an oscilloscope.

In addition to a shift in frequency, there will be a change in phase in pulse equipment. Here phase detection is used. The phase of successively received pulses is compared to a reference pulse in order to find any change in phase caused by the Doppler shift. There is no phase shift in echoes from stationary targets. By insuring can-

cellation of signals showing no phase shift, all evidence of stationary targets is removed; at the same time, echoes from moving targets are preserved.

## FORWARD TROPOSPHERIC AND IONOSPHERIC SCATTER COMMUNICATIONS SYSTEMS

### Ionosphere

The ionosphere is that region of the earth's atmosphere in which the gases are ionized by meteor activity and by ultraviolet light from the sun. Ions and electrons are distributed throughout the entire region, which is between 30 to 250 miles above the surface of the earth. Several layers have been formed by these ions and electrons in which the ionization density appears to have reached a maximum. These layers have been given the alphabetical symbols D, E, and F in their respective order of height, D being the lowest region. What is the quality of the ionosphere that gives it the properties of a mirror a great distance above the earth? What is it about the ionosphere that makes it return radio waves to earth, that permits long-distance radio communication? Really, the characteristics of the radio waves have as much to do with their being reflected as do the characteristics of the ionosphere.

The following definitions and facts concerning radio waves and the ionosphere will be helpful in understanding ionospheric scatter. The speed of a radio wave through space is about 186,000 miles or 300 million meters per second, the speed of light. The radio wave is made up of electric and magnetic fields, perpendicular to each other and to the direction of the wave. The ionosphere consists of ions and free electrons and is formed by meteors and by the bombardment of ultraviolet rays from the sun. The ionospheric gases have the characteristics of an electrical conductor, whereas ordinary air acts as an insulator.

*Reflection* is that phenomenon which causes a wave that strikes a medium of different characteristics to be turned back; the angles of incidence and of reflection are equal and lie in the same plane. *Refraction* is that phenomenon which causes a wave entering another medium obliquely to undergo an abrupt change in direction when the velocity of the wave in the second

medium is different from that in the first. *Diffraction* is the effect produced when waves (light, sound, or radio) encounter a barrier and bend around it. The mechanics of tropospheric and ionospheric scatter include reflection, refraction, and diffraction.

### Ionospheric Scatter

Until about 1950 it was considered impractical to use VHF communication for any distance greater than 200 miles. In early 1951, however, distances greater than 600 miles were found to be practical in the lower portion of the VHF band. These distances were obtained by a scattering process in the ionized E region of the atmosphere. Where operational requirements call for circuit paths from 600 to 1200 miles in length and for low susceptibility to long range jamming, ionospheric scatter type of communication has proven particularly useful. Since 1952 VHF ionospheric scatter communications systems have been used by the military to provide exceptionally reliable service in areas where high frequency and medium frequency systems are subject to severe propagational disruptions.

A simple analogy can be used to explain how ionospheric scatter is accomplished. Using a searchlight to illuminate a cloud layer at night, an observer can see a diffused reflection at distances well beyond line-of-sight. Most of the light passes through the cloud, but a small portion is scattered earthward. Substitute for the searchlight a VHF transmitter and directional antenna, for the cloud the ionized E region of the atmosphere, for the observer a receiver and another directional antenna (as shown in Figure 15-2), and you have the basic scatter system. Since the ionized E region is always present and dependable, it permits exceptionally reliable communication when the proper equipment is used.

There are two concepts used to help explain certain principles and techniques used in ionospheric scatter: scattering volume and common



Figure 15-2. Geometry of Ionospheric Scatter

scattering volume. Scattering volume is the volume in the ionospheric scattering region included by both the transmitting and the receiving beams, as illustrated in Figure 15-3.

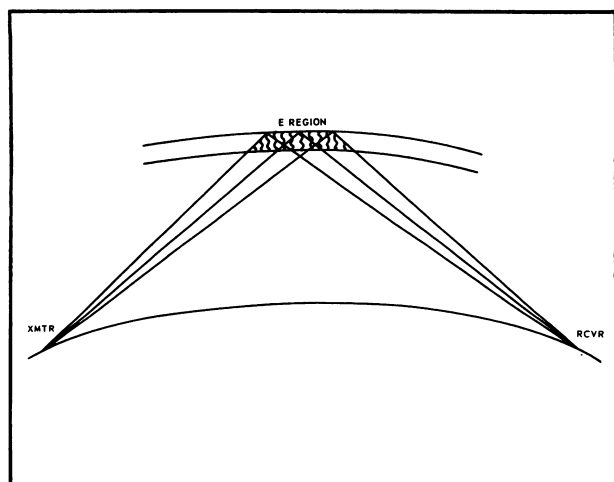


Figure 15-3. Scattering Volume

Common scattering volume is that region which is visible from both terminals, as illustrated in Figure 15-4. The surface path between the transmitter and the receiver must not be too long.

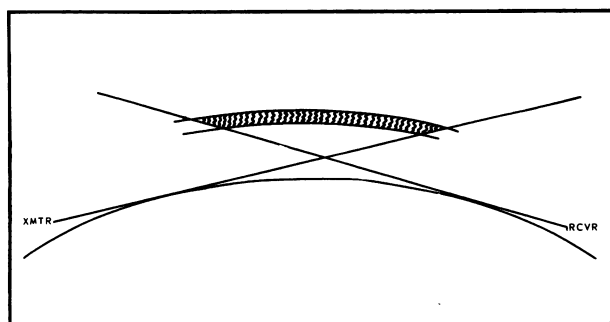


Figure 15-4. Common Scattering Volume

In summary, the scattering volume is that part of the common volume which is simultaneously within the beams of both antennas. The common scattering volume includes all of the ionospheric scattering region that can be simultaneously *seen* from both the transmitting and the receiving antennas.

## General Theory of Scattering

A number of theories have been proposed to explain the nature of ionospheric scatter. To the present time, however, no single theory has been universally accepted as a complete explanation. Each theory, while acceptable in most respects, has at least one disadvantage that prevents it from being completely acceptable.

One group of theories takes into consideration variations in the texture of the ionosphere. These may be due to acute variations in density with height (partial-reflection theory), or to discontinuities caused by turbulent fluctuations of ionization in the lower atmosphere. These theories state that energy in small amounts is reflected or scattered from the ionosphere to the receiving antenna. Although small, the amount of energy is greater than would be expected to be bent, or diffracted, around the earth's surface. The rapid fading of the received signal with time normally observed in ionospheric scatter can be explained well enough by the theory involving turbulence, but not so easily by the partial-reflection theory. In the latter theory, changes in density with time must be theorized to take place to account for the fading.

A different theory states that ionospheric scatter is largely the result of scattering from meteor trails. Meteors, traveling at high speeds, upon entering the earth's upper atmosphere generate heat that ionizes the atmospheric gases. The ionized gases will reflect radio energy. Rapid fading can be explained with this theory, depending as it does on the combined efforts of a large number of meteors. This theory can also be used to explain the variations in the direction of arrival of received energy, an observed characteristic of ionospheric scatter transmission.

It is most likely that the true method of ionospheric scatter propagation is actually a combination of the theories just described. All actions take place at the same time but with relative importance, changing with time of day and season in a more or less systematic manner. Whatever the final conclusion may be, there are certain recognizable characteristics that have to be considered. One is the effect of solar radiation. Signal intensities change with the time of day and seasons. This indicates that the effect the sun has on the ionosphere is an important factor.

Another fact that has to be considered is the effect of the meteors. The correlation of increased signal intensity and increased meteoric activity proves that ionospheric-scatter signals are affected by the passage of meteors. Other factors that must be included in the general theory are variable directions of signal arrival at the receiver, rapid fading, and corpuscular radiation, presumably of solar origin.

It is very interesting to note the effects of distance and received signal intensities. The turbulence theory indicates that the signal does not begin to reach a usable level until the path length exceeds 600 miles. From that point the received signal intensities increase until a broad maximum is reached at about 800 miles. After this distance the signal intensity gradually decreases. This discussion, of course, assumes the proper antenna height.

As the frequency is increased, there is a reduction in signal strength. Frequencies between 30 and 40 mc are the most efficient ones for ionospheric scatter. The polarization has little effect on received signal strength, although horizontal polarization is slightly better than vertical polarization. The signal at the receiving antenna has both horizontal and vertical components regardless of transmitted polarization.

### **System Capabilities**

Voice, teleprinter, and facsimile can be transmitted by ionospheric scatter propagation. The vulnerability of such circuits to long range jamming is very low. Continuity of service is exceptionally high because of the ability to operate continuously at a single frequency. Ionospheric scatter systems may be designed to have very high equipment reliabilities and hence low maintenance outage. These qualities are provided through dual operating equipment. This feature is important, as experience has shown that overall system reliability tends to be governed by equipment reliability rather than propagational difficulties.

The capability of the system to handle various transmission rates and types of information is normally provided by the use of conventional multiplexing and modulation techniques. Both time-division and frequency-division multiplexing methods may be used. Time-division multiplexing is the orderly transmission of signals of the same frequency by automatic electronic

switching. Frequency-division multiplexing is the orderly transmission of signals of different frequencies by automatic electronic switching.

**TRANSMITTING EQUIPMENT.** Transmitter power amplifiers used in ionospheric scatter installations have the following major characteristics: (1) output frequency of 30 to 60 mc, (2) output power of 20 to 60 KW, and (3) high power-conversion efficiency, on the order of 50%.

Careful attention should be given to linearity, to internally generated noise, and to shielding of strong r-f fields. It is also desirable to have an easy changeover from linear operation to Class C operation.

**RECEIVING EQUIPMENT.** Important characteristics of teleprinter channel receivers are: (1) low noise figure, 2 db for the high band (45 to 60 mc) and 3 db for the low band (30 to 45 mc); (2) provision for diversity reception; (3) narrow i-f bandwidths; and (4) frequency stability.

For voice receivers, a narrow-band FM or single-sideband receiver may be used. If voice signals are frequency-division multiplexed with teleprinter carriers, the receiver selectivity must be adequate to reject teleprinter channels that may be only a few kilocycles away.

**ANTENNA EQUIPMENT.** The performance of an ionospheric scatter system is directly related to the qualities of the antennas used. The most common are the rhombic array, the Yagi array, and the corner-reflector.

### **Forward Tropospheric Scatter**

A tropospheric scatter facility is capable of two-terminal communication over distances up to 300 miles. Since ionospheric scatter communication is usually not effective on paths under 400 miles, and since microwave links, because of terrain, are sometimes not practicable on long paths, forward tropospheric scatter communication fills a gap in the USAF communication system. To provide reliable communications on a path of 300 miles, high engineering standards must be met in both propagation and equipment. The propagation design may be controlled to fit the existing conditions; however, the equipment, once manufactured, is fixed.

The logistical and engineering requirements of a forward tropospheric scatter facility are little different from those of most radio systems. The scatter equipment requires 130 to 400 KVA of power, the amount dependent upon the type of

system that is chosen. As with similar facilities, high, level land areas are best suited for the installation and operation of this type of communications.

Each forward tropospheric scatter communications system must be individually designed, because the operating command has specific requirements. The most important consideration, usually decided by the operating command, is the reliability of the system. Two factors control the design of the system reliability. A breakdown of the equipment will lower the reliability of the service, but this condition may be minimized by providing dual equipment. The second factor, which is completely a design function, is propagation reliability.

The system is generally designed with enough flexibility to obtain any reasonable degree of reliability. Over a 300-mile path, reliability of 99.99% may be attained, although certain factors may limit the number of voice channels or require high power and large "dish" antennas. The important fact, however, is that this type of communications may be achieved. When the reliability requirement is lower, the number of voice channels may be increased, and the use of smaller antennas and lower power may prove sufficient. Reliability is a function of path distance and path conditions, then, to which all computations and design requirements are made to conform.

For nonstrategic voice transmission, 95% reliability is considered adequate; this lower reliability usually results in a saving of thousands of dollars. For a missile guidance system, 95% reliability could prove disastrous; therefore, such a system requires a higher reliability. The reliability is determined from the type of service required. Various bandwidths and numerous channels can be used. The forward tropospheric scatter terminal equipment is compatible with voice systems that terminate in telephone line pairs and use 600-ohm and 135-ohm line circuits for ringdown.

**GENERAL THEORY OF SCATTERING.** As with ionospheric scatter, many theories have been presented on the mechanics of tropospheric scatter. Some theories have been withdrawn because of new data and experimentation with this phenomenon. The theory presented here is probably the most widely used and also the most conservative presented to date. It has been compiled

from the latest and most reliable reports and articles.

The troposphere is the region of the earth's atmosphere that extends from the ground to a height of slightly over six miles. It lies several miles below the stratosphere and the ionosphere. The difference between the ionosphere and the troposphere is that there is practically no ionization of the air molecules in the troposphere. Also, there is a decrease in air pressure as you ascend from the earth's surface toward the stratosphere.

The ability of the troposphere to act as a smooth refractive medium is based upon the variation of the dielectric constant. The greater the decrease in dielectric constant, the greater the bending of the beam toward the earth. Water vapor content is the main cause of the variation of the dielectric constant. The greatest variations in heating and cooling take place near the surface of the earth; therefore, the greatest variations in the dielectric constant also occur near the surface of the earth. It can be said that the vapor variations play no great part above four miles, and have no serious effect on predicting the performance of tropospheric propagation.

As the earth's atmosphere is in a constant state of motion with respect to the earth, small irregularities, or eddies, occur. These irregularities occur in "blobs" that are large compared to the wavelength used in scatter communications; thus, they present a different index of refraction from that of the surrounding medium. The index of refraction is the ratio of the velocity of the wave in free space to that in the medium. The changes in the refractive index are a result of the variations in the dielectric constant. These variations, as we have said, are a result of the turbulent air motion and the water content.

This abrupt change in the index of refraction produces a "scattering" of an electromagnetic wave. However, most of the propagated energy continues in the forward direction, although enough energy is scattered toward the earth to be usable. Effectively, the blob has reradiated the signal toward the earth. The signals received will be affected by the actual conditions of the atmosphere, since scatter depends upon turbulences in the atmosphere. The number of blobs in the volume common to both the transmitting and the receiving antenna beams (Figure 15-5) determines the amplitude of the signal received. The scatter model in Figure 15-5 is idealized to



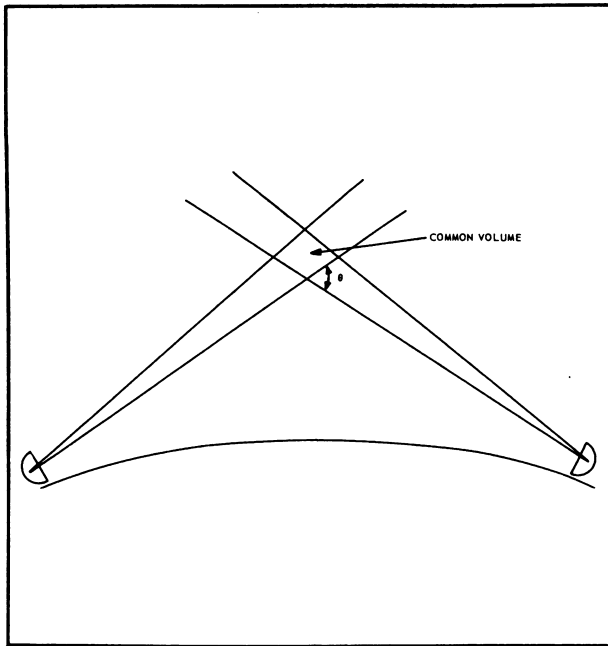


Figure 15-5. Tropospheric Scatter Geometry

indicate the path geometry. Notice that both the receiving and the transmitting antennas are directed at the horizon. A certain volume of the atmosphere is common to both antennas. It is within this volume that the scatter phenomenon takes place.

It can be proven mathematically that the scatter angle ( $\theta$  in Figure 15-5) is the most important factor in determining the received power.  $\theta$  is the angle between a ray from the transmitter antenna and one to the receiving antenna. If the scatter angle is increased, the received power falls off rapidly. With wide-beam antennas, the effective scatter volume is increased, but the effective received power is decreased. With narrow-beam antennas, the entire common volume contributes significantly to the received signal. For each ray of the entire beamwidth considered, a slightly different angle exists. It is necessary to sum up the infinite number of rays over the entire volume to obtain the sum of the individual contributions. It is correct to assume, therefore, that the scatter radiation comes from an infinite number of small scatterers within the common volume and that each scatterer effectively reradiates according to the amount of energy impinging upon it.

**SCATTER CAPABILITIES.** There is a great similarity between forward tropospheric scatter and

ionospheric scatter in that the upper atmosphere must have different refractive indexes for communication to be possible. Ionospheric scatter communications are most effective on paths exceeding 400 to 500 miles, and their propagation techniques are slightly different from those used in tropospheric scatter communications. While tropospheric scatter is efficient and reliable up to approximately 300 miles, ionospheric scatter is commonly successful on links exceeding 1000 miles. Reliability requirements usually dictate the total path length of either type. Tropospheric scatter is used when a large number of channels is required, because its shorter path lengths make it more efficient than ionospheric scatter. Ionospheric scatter has a limited total information bandwidth.

Tropospheric scatter may be called upon to transmit any type of signal, although voice transmission is usually used. Normally, some teletype will be transmitted, using equipment that feeds four teletype channels into one voice channel. With modifications to certain equipment, facsimile, telemetry, video, and other types of communications may be transmitted and received. Total path attenuation due to the scatter phenomena is dependent upon frequency, distance and horizon angle.

**ATMOSPHERIC CONSIDERATIONS.** The atmosphere that surrounds the earth is stratified, and all the strata have varying characteristics and relationships. Scattering blobs are actually relatively small volumes in space at which the index of refraction varies from the surrounding atmosphere. Usually they are caused by the meeting of moisture-laden air masses and warm air masses, which effects a change in refractive index. Rain, snow, and similar weather conditions do not have a persistently noticeable direct effect on VHF and UHF waves.

Cosmic noise has an effect on radio propagation. Cosmic noise encompasses the radio interference caused by eruptions from sun and stars. Eruptions cause an abnormal ionization of the strata affecting the reflection and refraction conditions necessary for propagation. It is impossible to predict accurately the nature of these disturbances or even the effect of the phenomena. But, at the frequencies used in tropospheric scatter, cosmic noise and cosmic interference are usually so small that they do not interrupt communications. Man-made noise at

the tropospheric scatter frequencies are usually of some consequence, however, and must be considered in the design and the location of scatter systems. Usually the least costly and simplest solution is to locate the system away from vehicular traffic, aircraft, machinery, electrified fences, power lines and other causes of man-made interference.

**ANTENNAS.** Two basic antenna systems are associated with Air Force tropospheric scatter systems. A wide variety of communication facilities is provided by the combination of these antenna systems with other equipment. It is possible to use either a 28-foot or a 60-foot parabolic reflector illuminated by either a vertically- or a horizontally-polarized feedhorn. If two reflectors are used, a vertically-polarized feedhorn to one reflector and a horizontally-polarized feedhorn to the other will provide a higher order of diversity. However, this necessitates cross-polarized feedhorns for reception at both terminals.

Due to the narrow beamwidth of the antennas, it is necessary that the equipment be made capable of changing both elevation and azimuth.

## **MOVING TARGET INDICATOR SYSTEM**

### **General**

The Moving Target Indicator (MTI) system effectively cancels clutter caused by fixed target signals, and thus reveals moving target signals. The unwanted echoes can consist of ground clutter (echoes from surrounding objects on the ground), sea clutter (echoes from the irregular surface of the sea), or even of echoes from storm clouds. The problem is to find the desired echo in the midst of the clutter. To do this, the MTI system must be able to distinguish between fixed and moving targets and then eliminate only the fixed targets. This is accomplished by phase detection and pulse-to-pulse comparison.

Target echo signals from fixed objects have a definite fixed phase relationship from one receiving period to the next, whereas moving objects produce echo signals that have a different phase relationship from one receiving period to the next. It is on this principle that the MTI system operates to discriminate between fixed and moving targets.

Elimination of fixed targets is obtained by delaying the signals received from each transmitted pulse for a period of time exactly equal to the pulse recurrence time, and then combining the delayed signals with the signals received from the next transmitted pulse in such a manner that the amplitudes subtract from each other. Since the fixed targets give the same amplitude on each successive pulse, they will be eliminated. But the moving target signals are, in general, of different amplitudes on each successive pulse so that the signals do not completely cancel but leave a small signal that can be amplified and presented on the indicators.

### **Block Diagram Analysis**

The over-all operation of the MTI system may be seen by an examination of the block diagram of the Signal Comparator Unit of a typical radar set shown in Figure 15-6. This unit contains those stages that are the difference, essentially, between a normal receiver and an MTI receiver.

As shown in Figure 15-6, 30-megacycle signals from the signal mixer are applied to the 30-megacycle amplifier, where they are amplified, limited, and fed to the phase detector. Another 30-megacycle signal, obtained from the coho (coherent oscillator) mixer, is applied as a lock pulse to the coherent oscillator. The coho lock pulse is originated by the transmitted pulse and is used to synchronize the coherent oscillator to a fixed phase relationship with the transmitted frequency at each transmitted pulse. The 30-megacycle c-w signal output of the coherent oscillator is applied, together with the 30-megacycle echo signal, to the phase detector.

The phase detector produces a video signal whose amplitude is determined by the phase difference between the coho reference signal and the i-f echo signals. This phase difference is the same as that between the actual transmitted pulse and its echo. The resultant video signal may be either positive or negative. This video output, called coherent video, is applied to the 14-megacycle carrier oscillator.

The 14-megacycle c-w carrier frequency is amplitude-modulated by the phase-detected coherent video. The modulated signal is amplified and applied to two channels. One channel delays the 14-megacycle signal for a period equal to the time between transmitted pulses. The signal



is then amplified and detected. The 2500-microsecond delay required (the period between transmitted pulses) is obtained by using a mercury delay line or a fused quartz delay line, which operates ultrasonically at 14 megacycles.

The signal to the other channel is amplified and detected with no delay introduced. This channel includes an attenuating network that introduces the same amount of attenuation as does the delay line in the delayed video channel. The resulting nondelayed video signal is combined in opposite polarity with the video signal resulting from the *preceding* echo signal, that is, the delayed signal.

The amplitude difference, if any, between the two video signals is amplified, and since the signal may be of either polarity, it is made unipolar. The resultant video signal, which represents only moving targets, is sent to the indicator system for display.

An analysis of the MTI system operation just described shows that signals from fixed targets, which have an unchanging phase relationship to their respective transmitted pulses, produce in the phase detector recurring video signals of the same amplitude and polarity. Thus, when one video pulse is combined with the preceding pulse of opposite polarity, the video signals cancel, and no information is passed on to the indicator system.

Signals from moving targets, however, will have a varying phase relationship to the transmitted pulse. As a result, the signals from adjacent receiving periods produce signals of different amplitudes in the phase detector. When such signals are combined, the difference in signal amplitude provides a video signal that is sent to the indicator system for display.

The timing circuits shown in Figure 15-6 are used to accurately control the transmitter pulse recurrence frequency to insure that the pulse repetition period remains constant from pulse to pulse. This is necessary, of course, for the pulses arriving at the comparison point to coincide in time and thus achieve cancellation of fixed targets.

As shown in Figure 15-6, a feedback loop is used from the output of the delay channel to the trigger generator and gating multivibrator circuits. The leading edge of the square wave produced by the detected carrier wave in the delayed video channel is differentiated at the

pick-off amplifier and used to activate the trigger generator and gating multivibrator. The trigger generator sends a trigger to the modulator, causing the set to transmit.

The gating multivibrator is also triggered by the negative spike from the differentiated square wave. This stage applies a 2400-microsecond negative gate to the carrier oscillator. Thus, the 14-megacycle oscillator operates for 2400 microseconds and then is cut off. Since the delay line time is 2500 microseconds, the 14-megacycle oscillations stop before the initial waves reach the end of the delay line. This wave train, when detected and differentiated, turns the gating multivibrator on, producing another 2400-microsecond wave train. The 100 microseconds of the delay period when no signals are applied to the delay line is necessary in order that the mechanical waves within the line have time to damp out before the next pulse repetition period.

In this manner the pulse repetition period of the set is controlled by the delay of the mercury or quartz delay line. Since this delay line is also common to the video pulses going to the comparison point, the delayed and the undelayed video pulses will arrive at exactly the same time.

## BASIC TELEVISION

### Introduction

The scope of this discussion is intended only to familiarize you with the underlying principles involved in the transmission and reception of television signals. By definition, television is the system of telecommunication for the transmission of transient images of fixed or moving objects. Television is a comprehensive subject. Almost every type of electronic circuit is involved in its transmission and reception. A few years ago the radio was the principal means of communication via air waves. Today, television is a vital factor in the home, and closed circuit TV is rapidly becoming a new approach to instruction in the school.

### Television Signal

There are certain requirements that must be met for the television signal to be satisfactory from an observer's point of view. The image should have the same smooth appearance as

that of a good photograph. There must be absolutely no flicker. The size of the picture should be such that it can be viewed by several people who are eight feet or more from the screen. There must be an adequate amount of light from the receiver screen to meet the changing requirements of day or night viewing. The *aspect ratio* must be such that the picture "looks right" to the viewer. The aspect ratio is the ratio of the width to the height of the picture. This has been standardized as 4 to 3, making the picture height three-fourths of its width.

The essentials of a basic, modern television system are shown in Figure 15-7. In A, two transmitters are shown—one for picture and one for sound. Both transmitted signals, however, are picked up by a single receiver (B).

In the picture transmitter, a lens focuses the scene to be televised onto the photosensitive plate of the camera tube. (This plate is called a mosaic and will be explained later.) The mosaic is scanned by a thin stream of electrons from the electron gun of the camera tube. This scanning of the mosaic releases a series of electrical impulses, which become the video signal. The picture does not consist of one electrical impulse but of a series of signal impulses, each of which represents a very small area of the scene. These impulses are amplified and used to modulate the carrier, so that it can be transmitted into space. For the electron beam in the receiver tube to be synchronized with the beam in the camera tube, synchronizing pulses are inserted.

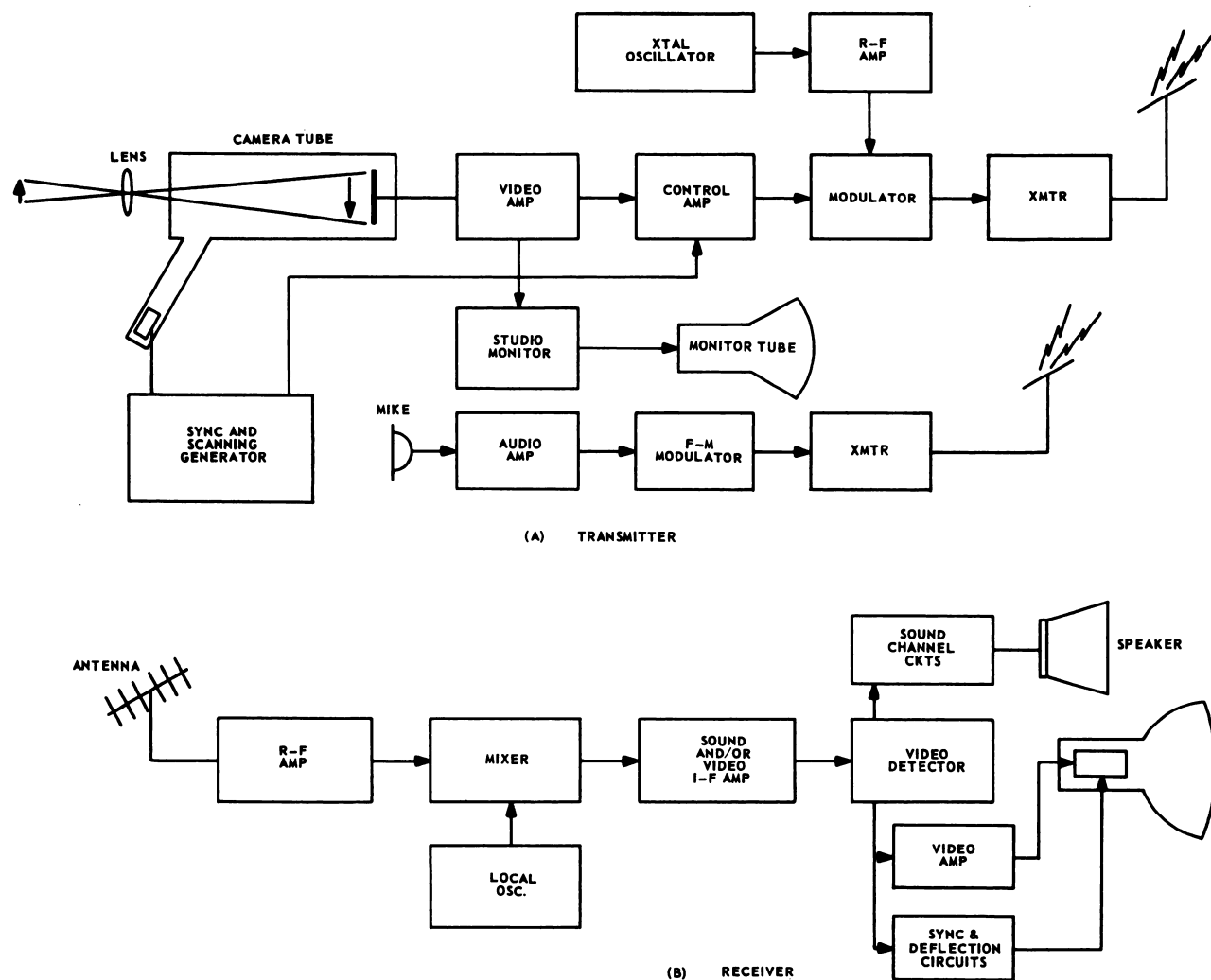


Figure 15-7. Simplified Basic Television System

The sound is transmitted by a conventional frequency modulation system, which consists of audio amplifiers, frequency modulators and a high frequency transmitter. Television is merely an extension of radio that includes both sound and picture, because in both cases, electromagnetic waves are used to transmit the signal.

As illustrated in Figure 15-7B, the video and sound signals are received by the same antenna and fed to the r-f amplifier. The diagram illustrates the sound being separated from the video after the mixer. The amplification of the sound signal at this point is less than that of the video signal. This is to prevent the sound signal from producing interference on the picture screen. After separation, the sound signal goes through one or more i-f amplifiers, an f-m detector, audio amplifiers, and the speaker.

After the mixer, the video signal is conveyed to the video i-f amplifiers, which produce two signal outputs. One is the video signal and blanking pulse. The other is the sync pulses that are fed to the sync and sweep circuits. There must always be a correlation between the sweeping electrons in the picture tube and the camera tube. The point at which the electrons strike the mosaic image must be the same point in the picture tube. This is the responsibility of the sync pulses. For black and white images, only vertical and horizontal sync pulses are required.

### Television Camera Tubes

The quality of image at the receiver is no better than that received and transmitted by the camera tube. There are several types in use. The one best suited to the purpose is used.

The *iconoscope* camera tube converts variations of light intensity in the picture elements into an orderly sequence of electrical impulses that correspond to the picture information. To do this, the photoelectric properties of the mosaic image plate are used.

It is known that certain metals have the characteristic of emitting electrons when light rays strike their surface. A light sensitive surface is usually prepared by evaporating a thin layer of caesium oxide on a silver base. A positive collecting plate is placed close to the emitting surface. The emitted electrons can be taken from the positive plate to produce an electric current that corresponds to the specific light ray.

The rectangular mosaic plate in the iconoscope is made up of many microscopic globules of sensitized caesium-silver compound that have been deposited on a thin sheet of insulating mica. The globules do not come into contact with each other, since they are insulated by the mica. On the opposite side, a continuous layer of a conducting substance is deposited; from this layer there is an electrical connection to an external circuit.

This construction actually forms a capacitor. The dielectric is the mica; the globule forms one plate, and the conducting substance on the backside of the mica the second plate.

The object to be televised is focused on the front face of the mosaic plate. Each globule emits electrons according to the light ray intensity reaching it; thus, each globule will assume a different positive charge, and will retain this charge, since it is insulated from the other globules. The mica also prevents any leakage of the charge to the conducting layer. The conversion of light variations into electrical impulses has now been accomplished. Figure 15-8 illustrates the construction of the iconoscope. The electron gun and deflection coils are similar in purpose to those of other CRT's. The electrical impulses are amplified and fed to the next stage.

The iconoscope will produce a good picture if the scene is sufficiently illuminated, but it has a tendency to produce background shading due to the secondary emission of electrons from the mosaic. A camera tube that is more sensitive than the iconoscope is the *image orthicon* (see Figure 15-9). The greater sensitivity of the image orthicon tube gives it two important advantages. One is the ability to televise scenes that have insufficient illumination to produce an acceptable image with other camera tubes. The other is the greater depth of field, permitting the inclusion of background that would otherwise appear blurred or obscured on the receiver screen.

In operation, the light rays from the scene to be televised are focused on a transparent photosensitive plate by a lens. From each point on the inner surface of this plate, electrons are emitted in proportion to the intensity of the incident light. The emitted electron image (the electron density at each point corresponds to the light at that point) is drawn to the target by a positive



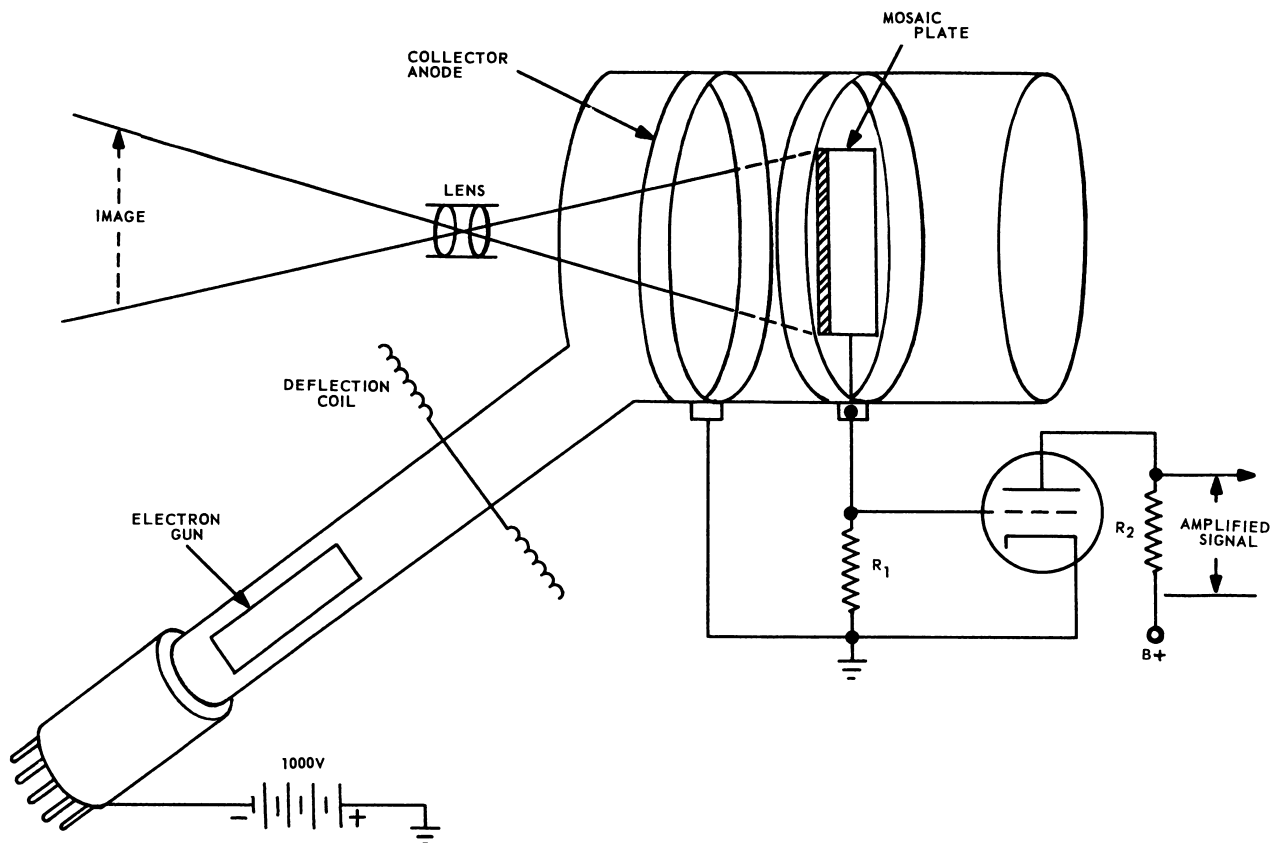


Figure 15-8. Internal Construction of Iconoscope

wall coating. At the target the accelerated electrons produce secondary emission, leaving a pattern of positive charges on the target due to the deficiency of electrons. This pattern corresponds to the pattern of light from the scene, with the most positive parts of the target representing the lightest parts of the scene.

At the same time the pattern is being formed on the target plate, the plate is being scanned by the beam from the electron gun. The velocity of the electron beam, as it approaches the target, is such that it is deflected and returns to the source. However, when the scanning beam moves over a section of the target containing a positive

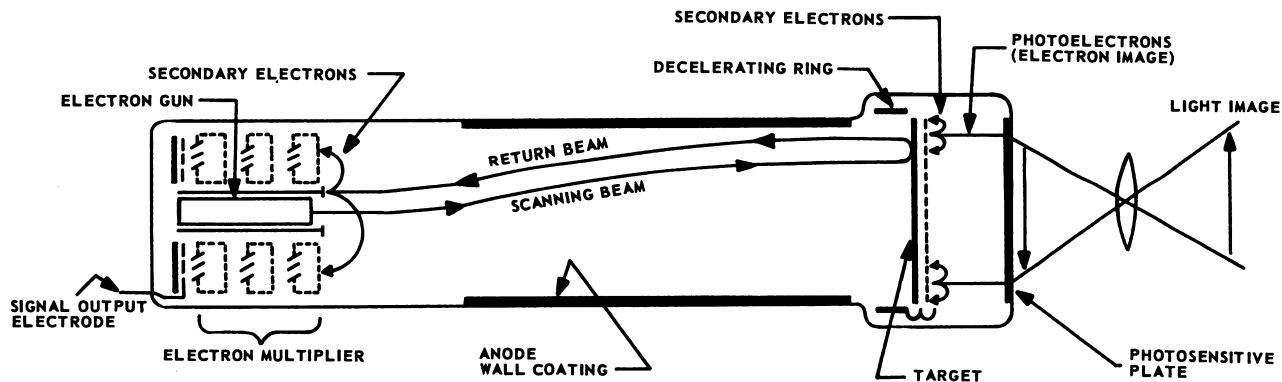


Figure 15-9. Construction of Image Orthicon Tube

charge, electrons are deposited to neutralize the positive charge. The returning electron beam, then, contains variations in magnitude that correspond to the charge distribution on the target plate and the picture elements in the optical image. Hence, the most positive points on the plate return the least number of electrons from the original scanning beam, and the voltage across the output resistor is inversely proportional to the positive charge intensity on the target.

The added sensitivity of the image orthicon is a result of several factors. The increased efficiency of the storage action on the target allows storage over a greater portion of the frame period. The combination of the higher photosensitivity obtainable with a conducting photocathode surface and the secondary-emission gain of the electron image at the target provides an increase of camera sensitivity. By the use of a two-sided target, the charge pattern can be formed on one side while the opposite side is scanned by the low-velocity electron beam. This reduces shading difficulties and provides stable operation over a wide range of scene brightness. An additional gain is provided by the electron-multiplier arrangement where a number of electrons are produced for each electron impinging thereon.

The image orthicon is used extensively in field television operations outside the studio, such as on-the-spot reporting of news, sports events, and conventions. Any scene with sufficient illumination for direct viewing by an audience can be

televised satisfactorily with the image orthicon camera because of its sensitivity and its ability to operate over a wide range of light values. In addition, the optical system used with the image orthicon is sufficiently compact to make it suitable for portable field equipment. While the image orthicon is also used for studio cameras, the iconoscope and image dissector tubes are preferred in many instances—particularly when televising motion-picture film.

The *image dissector* tube is considered to be an instantaneous scanner because the principle of storage is not involved. The tube, shown in Figure 15-10, contains a flat photosensitive cathode located at one end of the tube. The scene to be televised is focused on this cathode by a suitable lens system located outside the opposite end of the tube. As the light reaches the cathode, electrons are emitted in direct proportion to the amount of light striking any one point. Thus, the electron distribution leaving the cathode is the same as the distribution of light in the optical image. This electron image extends from the cathode surface and is attracted toward the aperture of the electron multiplier by the voltage on the nickel anode coating. As the electrons pass through the aperture, they are multiplied by secondary emission in the multiplier.

Since the aperture that receives the electrons (to provide the signal current) is stationary, the entire electron image must be moved past the opening in such a way that electrons are taken in the required succession of picture elements and

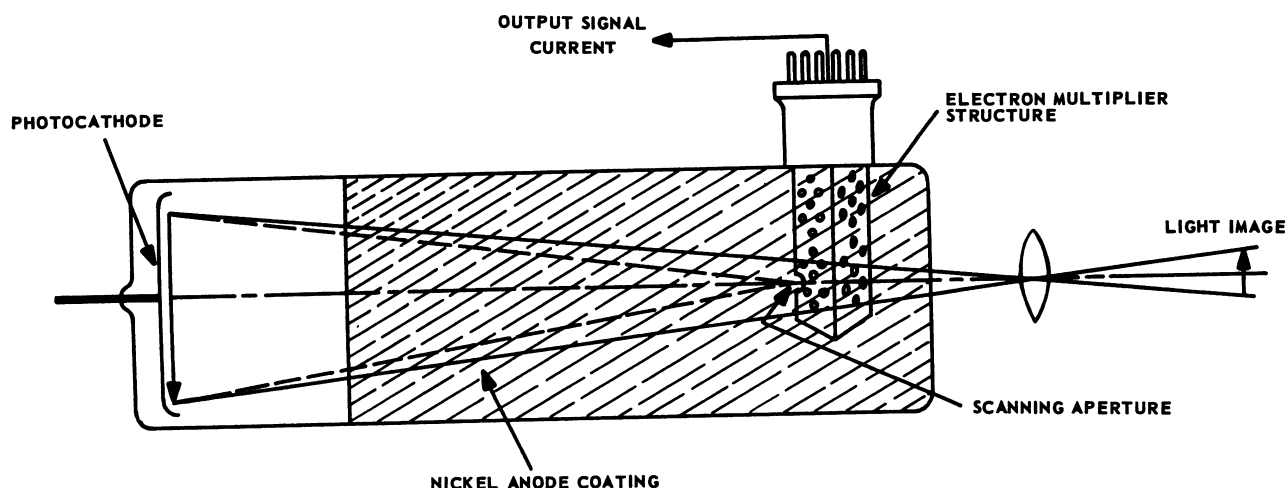


Figure 15-10. Construction of Image Dissector Tube

scanning lines. The electron image is moved past the aperture in the desired scanning pattern by vertical and horizontal deflecting coils mounted externally around the tube. The electrons that are not being scanned are attracted to the anode coating, but since the cathode provides a steady stream of electrons, the image is not destroyed.

The vidicon camera tube is similar to the image orthicon. The image is focused optically on a photoconductive target. When light strikes the target area, the scanning beam neutralizes the charged elements and current flows through the load resistor. The vidicon is physically smaller than the orthicon and is portable. Therefore, it finds wide application in industrial and tactical television systems. In fact, the portability of this tube has permitted the development of a man-carried television transmitter to be used for observation in forward combat areas.

### Television Scanning

In transmitting a picture, it is possible to transmit at one time all elements that compose the picture, or to transmit each element separately in a specific order. The complexity of the first system has been a determining factor in adopting the system of transmitting each element separately. Other methods of scanning that might be employed are narrow horizontal strips, narrow vertical strips, and spiral scanning.

To explain television scanning, let us review the scanning process of the conventional A-scan CRT. A horizontal sawtooth is applied to the horizontal deflection plates or coils and this causes the electron beam to travel from left to right across the face of the tube. On the negative-going portion of the sawtooth, the electron beam is retraced sharply back to the left side. Therefore, with no signal applied to the vertical deflection plates, a straight line appears on the screen.

Now apply a sawtooth voltage of one frequency to the vertical plates and a sawtooth voltage of a much higher frequency to the horizontal plates. The resultant is an electron beam traveling from left to right on the screen and also moving from top to bottom, as shown in Figure 15-11. A vertical sawtooth frequency of 60 cycles per second and a horizontal sawtooth of 15,750 cycles per second produced the lines shown. It is evident that the speed of the horizontal sawtooth is much greater than that of the vertical sawtooth.

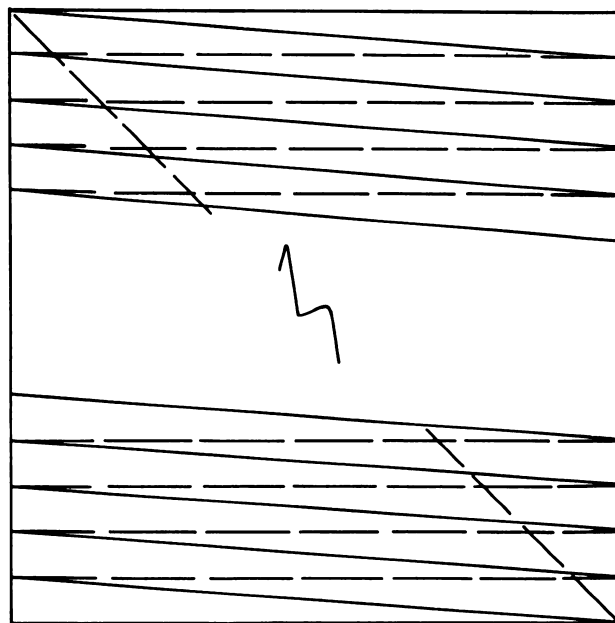


Figure 15-11. Basic Television Scanning

In fact, each line appears as an almost straight horizontal line. There are approximately 525 lines present from top to bottom of the television screen.

Let us now apply the scanning process to the iconoscope camera tube (Figure 15-8), and visualize the resulting action. Assume that the camera is focused on the scene to be televised and that each globule on the mosaic signal plate has a positive charge proportional to the light sensitivity it received.

At the first instant the electron beam, focused in the neck of the tube, is accelerated toward the upper left-hand corner of the mosaic signal plate. There, under the influence of the vertical and horizontal sawtooth voltages, it moves to the right and downward. As the electron beam moves to the right, it passes over the charged globules that are across the top of the mosaic plate. As each globule is scanned, electrons from the electron beam restore those electrons that have been emitted. This brings each globule back to its neutral state.

During the time that each globule is being neutralized, the charge that is held on the conducting surface of the opposite side of the mosaic is released. The releasing of this charge causes a small current to flow in R1 of Figure 15-8. The strength of the current is proportional to the charge on the neutralized positive globule which,

in turn, was proportional to the intensity of light focused on the mosaic at that particular point. In this manner, then, light rays have been transformed into equivalent electrical currents.

### Interlaced Scanning

In the modern television system, 525 lines constitute the scanning *raster*. By definition, raster is the predetermined pattern of scanning lines which provides substantially uniform coverage of an area. Since 15,750 lines are scanned per second, 30 frames are scanned per second. By definition, a frame constitutes the entire picture or image. The fundamental rate of 30 frames (images) was chosen because of its relation to the commercial power frequency. This frequency rate requires less filtering of the AC ripple, which causes the hum in the audio system.

In order to eliminate all possibility of flicker, the raster is scanned at an effective rate of 60 frames per second. To do this, the downward travel rate of the scan is increased so that every other line is scanned. This pattern is shown in Figure 15-12. The odd-numbered lines are scanned first, and then the beam returns and scans the even-numbered lines. As the odd and

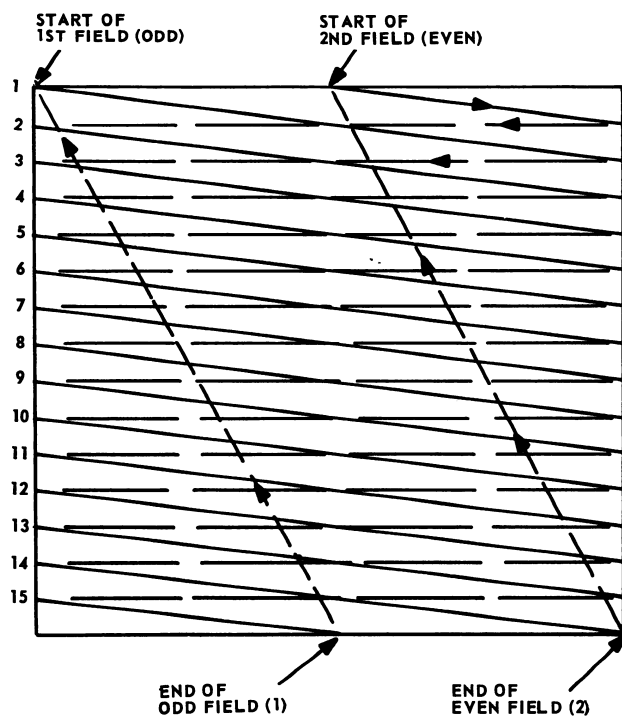


Figure 15-12. Interlaced Scanning

even lines are transmitted in  $\frac{1}{60}$  of a second each, both add up to  $\frac{1}{30}$  of a second. The scanning of both lines takes  $\frac{1}{30}$  of a second, so 30 frames per second is still the fundamental rate. But the *effective* rate is 60 frames per second and there is no flicker. This method of transmitting television images is called interlaced scanning.

### Compound Television Signal

The television signal is composed of three types of information. *Video signals* are the orderly sequence of pulses that represent the light intensity of the televised scene. The *sync pulses*, both horizontal and vertical, are rectangular pulses for locking-in both camera tube and picture tube. The *blanking pulses*, horizontal and vertical — are rectangular pulses for blanking out the screen during retrace time.

Another group of signals is transmitted to insure proper spacing of the interlaced scanning lines and to prevent loss of synchronization of the horizontal circuits during interlaced scanning. These are called *equalizing pulses*.

### Picture Signal

Figure 15-13 illustrates the entire transmitted signal. It depicts, from left to right, the last four lines being scanned, the vertical blanking time, and the first three lines of the next frame. During the vertical blanking time, the vertical pulse intervals hold the picture stationary vertically. These pulses of very short duration insure equidistant spacing between scanning lines. The equalizing pulses also permit transmission of identical vertical intervals between frames and between fields, and maintain horizontal synchronism during vertical trace intervals.

Figure 15-14 expands that portion of the video signal which depicts scanned lines. Notice that of the total amplitude available, approximately 80% is used for video signal variations. Where the signal voltage stops, the blanking voltage begins. Notice also that the blanking and sync voltages always have the same amplitudes, regardless of signal voltage, because of requirements in the television receiver.

Figure 15-14 also illustrates the polarity of signal as used in receivers in the United States. The reverse is used in England. The type used in the United States is designated as negative picture transmission, and the picture is said to

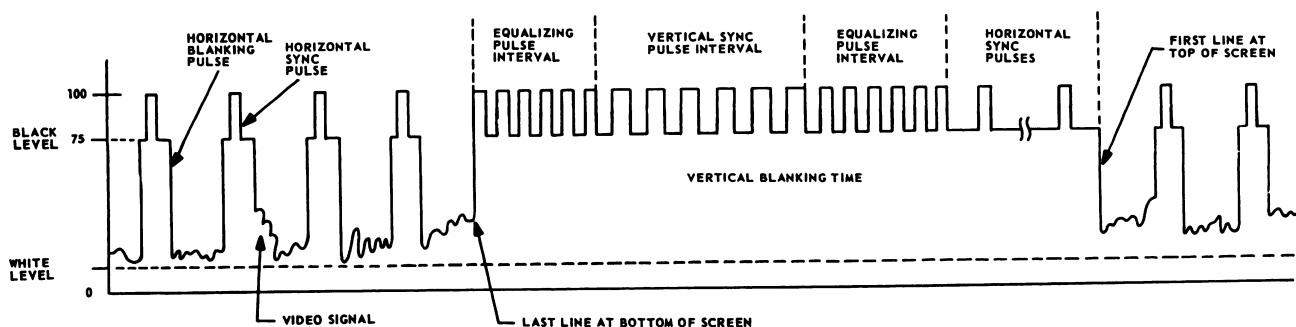


Figure 15-13. Complete Video Signal

be in the negative picture phase. The signal voltages from the iconoscope have been completely reversed. In the receiver the voltages must be changed again before being applied to the cathode of the picture tube. It is said that less interference and better over-all reception is obtained under this negative picture transmission.

### Signal Standards and Frequencies

Commercial television channels are 6 megacycles wide. In contrast, the entire radio broadcast band is slightly more than 1 megacycle wide; each broadcast channel is only 10 kilocycles wide. It is necessary to have a wide television channel in order to transmit pictures that have clarity and sharpness. Within the 6-megacycle band are found both picture and sound carriers and corresponding sidebands. Picture carrier and sidebands occupy approximately 5.75 megacycles, a very large percentage of the total band. The picture carrier is amplitude modulated, and the sound carrier frequency modulated.

### Vestigial Sideband Transmission

An unsymmetrical distribution of the frequency band of the picture carrier allows transmission with much better definition. This is called *vestigial sideband transmission*, with the high-frequency sideband being 4.5 mc wide and the low-frequency sideband being only 1.25 mc wide. Since the modern television station broadcasts signal components up to 4 megacycles wide, a 9-megacycle channel would be required to transmit a symmetrical sideband. In a 6-megacycle channel, leaving a frequency for the sound carrier and transmitting with symmetrical sidebands, the highest frequency picture component

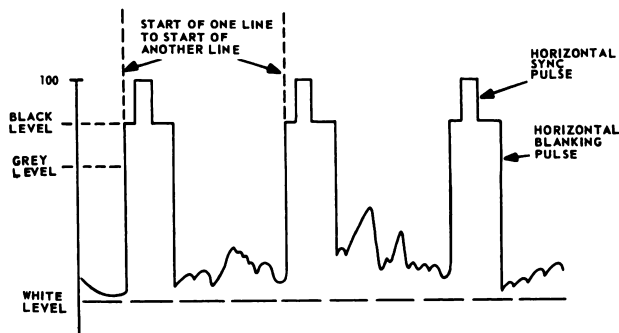


Figure 15-14. Expanded Video Signal

is a little over 2.75 megacycles. Picture components would be lost in transmitting just these frequencies. However, by partially suppressing one sideband, it is possible to transmit a 4-megacycle component in a 6-megacycle band. Figure 15-15 illustrates the unsymmetrical distribution in vestigial sideband transmission.

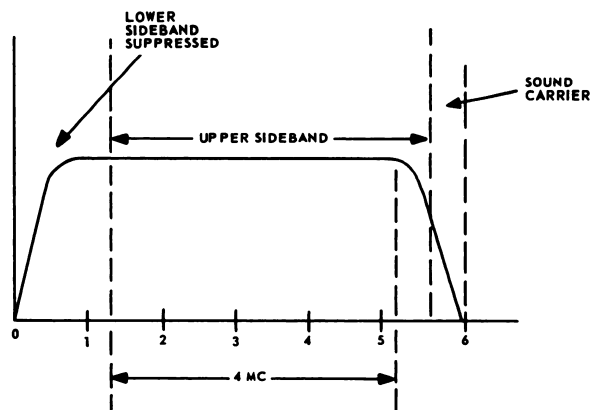


Figure 15-15. Frequency Distribution in Vestigial Sideband Transmission

The image to be telecast is focused on the image plate of the camera tube, which is comparable to the mike of the audio transmission. The iconoscope and the image orthicon are the tubes most generally used: the iconoscope for film projection and studio, the image orthicon for remote telecasts. Each tube has characteristics that make it desirable for its particular use.

The current resulting from the scanning of the image plate is amplified by a stage that is normally a part of the camera unit. The signal then goes to the distribution and control amplifiers, where the signal is mixed with signals from other cameras. Here the signal can be switched in and out or superimposed on another. Blanking and sync pulses are added at this time, being inserted with the proper polarity in respect to the black and white polarity of the picture signal. Let it be emphasized at this point that the sync generator is the timing unit of the entire system. In addition to acting as the basic timing unit, the sync generator initiates, shapes, and arranges in the correct sequence all the pulses required for control of the 525-line interlaced scanning.

After leaving the control amplifiers, the composite television signal goes to the modulator. The modulator has a unique assignment. It must amplify a single-polarity signal which is varying with the average brightness of the televised scene. The modulator must maintain a constant level of the blanking and sync pulses irrespective of the variations of the picture content and average brightness. The modulator is composed of DC amplifiers and/or DC restorers which maintain the level of the composite television signal.

The signal, after leaving the modulator, is fed to the transmitter unit, which is comprised of the frequency determining device, multipliers, amplifiers, and transmitter tube. The transmitter unit is comparable to any high-frequency transmitter.

## TELEVISION RECEIVERS

We have stated previously that the television receiver uses the same antenna to pick up the picture and sound signals. Therefore, it is necessary for both the antenna and the r-f section to have a response over the 6-megacycle band of frequencies that is uniform, yet selective enough to discriminate against undesired frequencies. In

a fringe area the antenna should be directional toward the transmitter site and should have a broad, flat bandpass.

In one early type of receiver, the video and sound signals are separated at the mixer and fed to their respective i-f amplifiers. The signals are then treated separately, the sound going to an FM detector and the video going to an AM detector.

In the *intercarrier-type* receiver, the video and sound signals are not separated until after they pass through a common i-f amplifier designed to yield a sound carrier with an amplitude of only about 5% of the maximum picture carrier amplitude. In the following video detector, heterodyne action produces a 4.5 mc sound i-f and all of the video frequencies. Even after passing through the video amplifier, the 4.5 mc sound i-f is low enough not to affect the picture, and must be amplified before going to the FM detector and audio amplifier.

## Controls

Figure 15-7B showed us the block diagram of a conventional-type television receiver, simplified. Before analyzing the circuitry of the receiver, let us briefly discuss the panel controls.

**CONTRAST.** This is the most important control on the front panel. This control adjustment is as necessary to the picture as the volume control is to the sound. The higher the video voltage applied to the control grid of the CRT, the more intense the image on the screen.

**BRILLIANCY CONTROL.** The brilliancy (brightness) control, usually tied to the cathode of the CRT, adjusts the DC bias on the picture tube, and is a working partner with the contrast control. The brilliancy control should not be adjusted until the contrast has been adjusted to the desired intensity of image. A proper setting of this control just biases the CRT to the point where the blanking voltage level cuts off the electron beam.

**FINE TUNING.** This control is a variable capacitor across the oscillator-mixer circuit. This control is important in that at high frequencies, small variations mean large frequency changes. The frequency desired is selected by either push button or switch, and the fine tuning compensates for small variations in the oscillator tuning circuit to center the signal on the i-f bandpass.



**HOLD CONTROL.** There are two hold controls, one for the vertical circuit and one for the horizontal circuit. If the picture should get out of sync, these controls enable the viewer to bring the oscillators within the range of control of the sync pulses.

Other variable controls, generally placed in the rear of the set, include horizontal linearity, vertical linearity, width control, height control, vertical centering, horizontal centering, and focus. There may be additional controls, depending on the complexity of the receiver.

### General Theory of Operation

Using Figure 15-7 for reference, let us now explain in a general sense the operation of the various blocks. The r-f amplifier has a three-fold function. First, it provides amplification of signals where the voltage value of the signal is the lowest. The r-f amplifier must operate linearly over a 6-megacycle band of frequencies for each channel. Gain is sacrificed to obtain amplification that is linear over a wide bandpass. Figure 15-16 illustrates a typical r-f amplifier for a TV circuit. Note the loading resistors across the tuned circuits.

Second, the r-f amplifier helps to discriminate against close frequencies in adjacent channels. Especially is this true for image frequencies. Tube interference, snow on the screen, is also lessened with proper circuit design. Third, local

oscillator radiation can interfere with other receivers that are nearby.

Detection in television is no different from that in radio; the intelligence (picture modulation) is removed from the i-f carrier. The modulation contains picture, blanking, and sync information. This is the composite television signal. As with any diode operation, there is no current flow when the cathode is more positive than the plate. The effect of this action, then, is a unipolar signal with the circuit designed to develop a signal of the proper polarity for use on the grid of the picture tube. At the same time the circuit must be designed to detect the highest and lowest frequencies of the i-f spectrum.

Since the American receivers use negative picture transmission, a positive output from the video detector is designated as negative picture phase. The signal is inverted before being applied to the grid of the picture tube. The simplified circuit in Figure 15-17 is used to show the inversion of signal for proper application to the grid of the picture tube.

The i-f system of a television receiver is much more complex than that of a radio receiver. Picture and sound must be amplified and separated. Figure 15-18 is a simplified drawing illustrating the separation of picture and sound. Gain over a band of frequencies 4 to 5 megacycles wide must be linearly amplified. Since the selectivity of the i-f stages is a factor that must be carefully

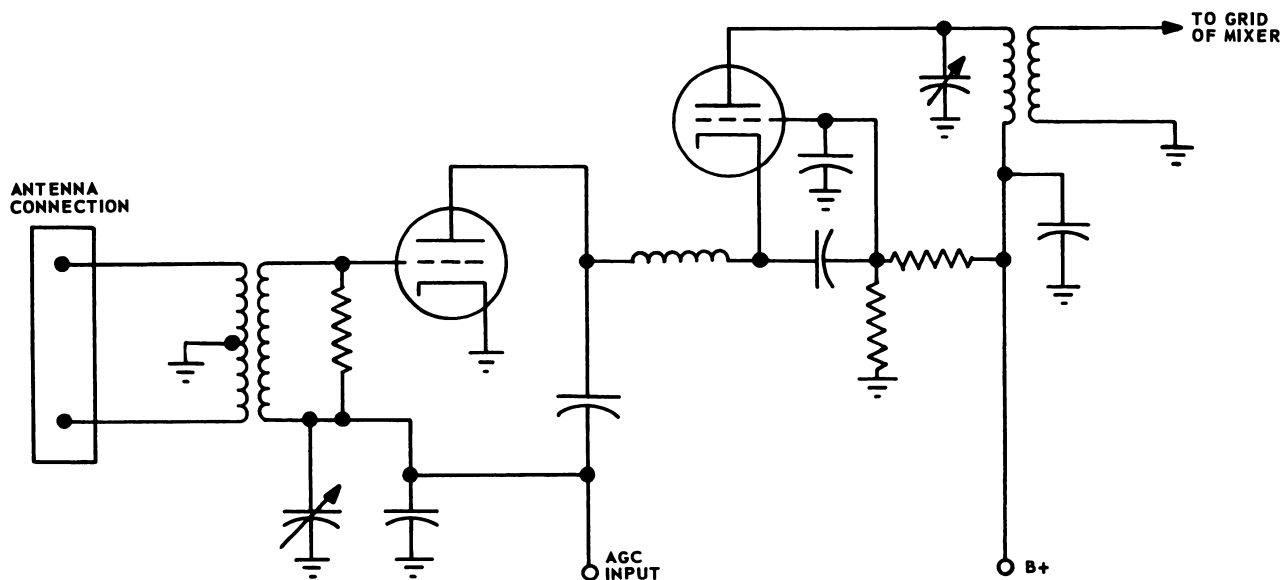


Figure 15-16. Typical R-F Amplifier

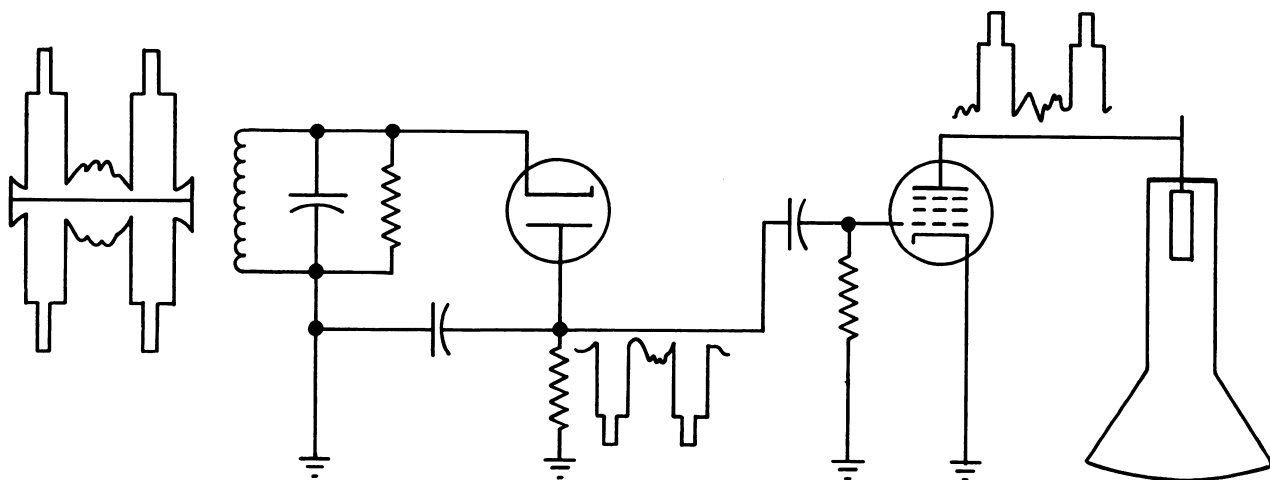


Figure 15-17. Detector, Inverter Amplifier Stages

considered, stagger tuning is used extensively. In fact, the majority of television sets of today use single-tuned coils between the stages of video i-f amplifiers. Each coil is tuned to a different frequency that is within the 4-megacycle bandpass of the system. The desired over-all response of all stages is a broad, flat bandpass. Unwanted frequencies that are present are rejected with wavetraps. The wavetraps are specifically designed to reject sound i-f frequencies that cause undesirable bar patterns on the picture tube screen.

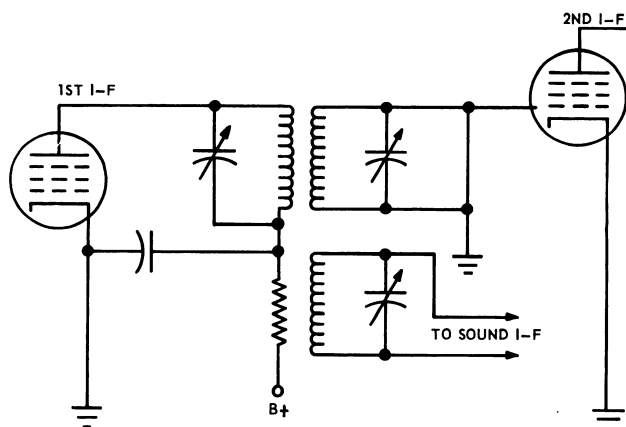


Figure 15-18. Picture-Sound Separation, Simplified

Many receivers use an Automatic Gain Control (AGC) system to prevent the undesirable effects of variations in signal strength. AGC

provides a voltage that varies the gain of the i-f amplifiers and keeps constant the peak amplitude of the signal applied to the video detector.

So far the discussion has taken the received TV signal through an r-f amplifier, converted it to an i-f frequency, further amplified it with i-f amplifiers, and detected it. More amplification is now necessary because the amplitude of the signal from the detector is not sufficient to provide the required contrast on the CRT.

While each i-f amplifier stage is required to pass a 4-mc band, the video amplifier must be designed to have a flat response over a band that extends from 30 cycles to 4 megacycles. The rectified video signal contains the picture, and blanking and synchronizing information, which must be passed without attenuation or phase distortion. Associated with the video amplifier is a DC restorer. The purpose of the DC restorer is to return all blanking voltages to the same level as they were in the incoming signal. The shift in the blanking level is caused by the charge and discharge characteristics of any RC coupling between stages. If the shift of the blanking level is not corrected, we will have either a picture that is too dark or too bright.

The composite video signal shown in Figure 15-13 includes the synchronizing pulses that are required to synchronize transmitter and receiver scanning. The sync pulse is applied during the blanking period to prevent it from interfering with the picture information contained in the camera signal. The horizontal sync pulse at the

end of each line causes the beam to return to the left side of the screen in position for the next line. The vertical sync pulse at the end of each field causes the beam to return to the top of the image in position to begin the next field.

Note that the amplitude of the signal is divided into two sections; the lower 75% is devoted to the active camera signal, and the upper 25% is used for the sync pulses. The smallest amplitudes of the signal represent the whitest parts of the picture, while the higher amplitudes represent the darker picture information until the black level is reached. The black level is maintained at a constant percentage of the peak signal amplitude in order to maintain a brightness level in the television system.

The combination of blanking pulses, sync pulses, and image signal is shown in Figure 15-13 in which four complete horizontal lines of the picture are represented. At the end of each line, the blanking pulse is applied to the beam. This prevents the electrons from reaching the mosaic of the camera tube or the screen of the receiver. While the blanking signal is being applied, the sync pulse causes the horizontal deflection coils to move the position of the electron beam from the right side of the picture to the left. When the blanking voltage is removed, the electron beam starts scanning again. This process continues until all the lines in one field have been scanned. The time between the beginning of the blanking pulse and the beginning of the sync pulse is called the *front porch*, while the time between the end of the sync pulse and the end of the blanking pulse is called the *back porch*.

The vertical motion stops at the bottom of the field. Therefore, it is necessary to bring the beam quickly to the top of the picture so that the next field can be scanned. As the vertical triggering and retrace require a longer period of time than the horizontal triggering and retrace, a longer blanking signal must be used.

The left side of Figure 15-13 shows the last four lines at the bottom of a field with the required horizontal blanking and synchronizing pulses. After the last active line, the signal is brought up to the black level by the vertical blanking pulse in preparation for the vertical retrace. To provide exact timing of the vertical retrace motion in successive fields, the vertical blanking period begins with six equalizing pulses

spaced at half-line intervals and ends with six equalizing pulses.

The vertical sync pulse is broken up to permit the horizontal sync voltages to continue without interruption. Otherwise, the horizontal oscillator that controls the retrace in the receiver would slip out of control with a resulting loss of synchronization. The serrated pulse permits both retraces to occur at the proper time. One vertical pulse occurs at the end of every  $262\frac{1}{2}$  lines, while a horizontal pulse is applied at the end of each line.

## TELEVISION ANTENNAS

### Introduction

The antenna system is the link between individual transmitting stations and the multitude of widely scattered receivers. Regardless of the form of the television antenna, its main purpose is to radiate or receive signal-bearing r-f energy. The considerations involved in the design of a television receiving antenna are somewhat different from those of the broadcast receiving antenna. First, the television antenna must obtain a good signal over an extremely wide range of frequencies. Second, the amount of signal voltage that can be induced is limited by the short lengths of television antennas. Third, the problem of obtaining adequate signal is increased by the low signal strengths encountered in VHF. The problems that are particularly important in the television antenna system include selecting and positioning of a suitable antenna, selecting and installing a suitable transmission line, and proper matching of antenna, transmission line, and receiver impedances. Each of these items is critically important in receiving good pictures.

### Propagation of Direct, Sky, and Ground Waves

In considering propagation in the 30- to 1000-megacycle band, which contains the television channels, only the direct-wave mode of propagation is effective at all times. The direct wave, also called the space wave, is defined as that part of the radiated energy which travels directly from the transmitting antenna to the receiving antenna. No portion of the direct wave is returned from the ionosphere above or reflected from the surface of the earth or objects on or above it.

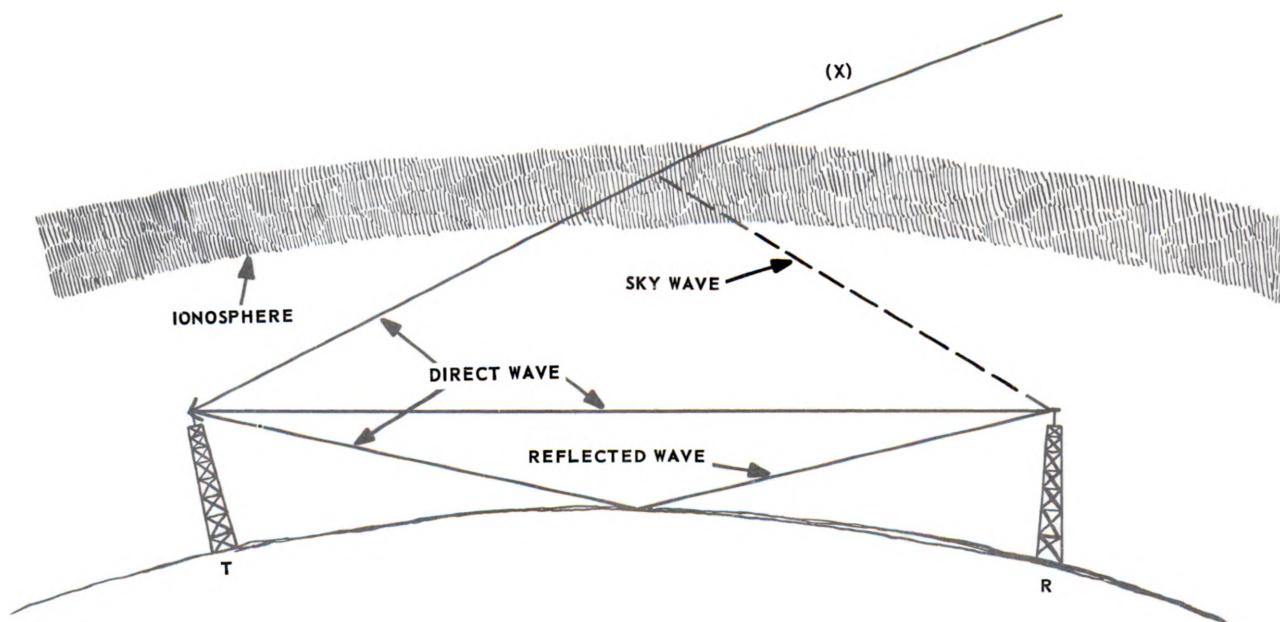


Figure 15-19. Direct, Sky, and Reflected Waves

Sky-wave propagation and ground-wave propagation, which account for the long distances covered by radio waves at lower frequencies, become less effective as the frequency is raised above 30 mc. At frequencies of 100 mc and above, they are of little practical use except under unusual circumstances.

Figure 15-19 shows the portions of the radiated wave in the sky wave, the reflected wave, and the direct wave. The sky wave is considered to be that portion of the radiated wave (shown by the dotted line) that reaches the receiver antenna after being deflected downward from its original direction. This wave becomes very weak when the frequency is increased. At 30 mc and above, most of the radiated energy passes through the ionosphere with only slight downward bending, excepting under unusual ionospheric conditions.

Ground wave propagation refers to the part of the radiated energy that moves along the surface of the earth at and closely above the boundary between the surface and the air above it. At frequencies below 3 mc, the ground wave (sometimes called the surface wave) is strong enough to permit communication at distances far beyond the optical horizon. At the higher frequencies, however, the attenuation of the ground wave is great, and the true ground or surface wave decreases in strength so rapidly that it is useless

for practical purposes. Because of these conditions, sky and ground waves are relatively unimportant in television transmission most of the time.

The radio energy intercepted by a receiving antenna located relatively near but above the surface of the earth is actually the resultant of the direct and reflected waves which reach that point. As a result, the actual strength of the signal generated in the receiving antenna may be either stronger or weaker than that which would be produced by the direct wave alone; the strength of the signal depends on the phase relationship between the direct and reflected waves at the antenna. The reflected wave is usually weaker because of some absorption and scatterings, and the two waves seldom will cancel each other completely, even when they are exactly out of phase at the receiving antenna.

Because of the relative ineffectiveness of sky-wave and ground-wave propagation in the VHF and UHF bands, only those points on or above the surface of the earth that can be reached by the direct wave are normally within effective communication range. The general effect on communication can be seen in Figure 15-20. Antennas located at B and C receive good signals, but an antenna at D receives only a very weak ground-wave signal.



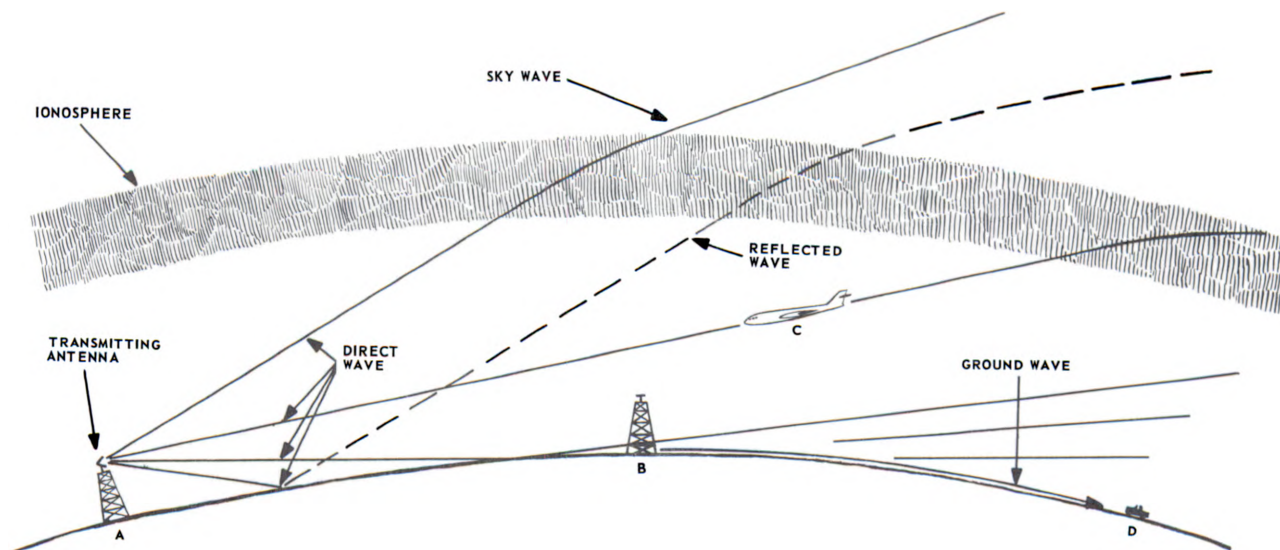


Figure 15-20. Practical Propagation at VHF

Radio waves above 30 mc are often referred to as quasioptical to indicate that they behave similarly to light waves. Since radio waves in this frequency range behave similarly to light waves, their propagation characteristics are often called line-of-sight transmission and reception. Because electromagnetic waves move in straight lines in free space, it might seem at first that communication beyond the range at which the receiving antenna actually can be seen from the transmitting antenna would be impossible. Experiment has shown that this is not true.

The radio horizon is somewhat beyond the optical horizon, even under normal conditions of the troposphere (lower part of the atmosphere). The radius of the troposphere considered with the earth is 1.33 times as great as the actual radius of the earth alone. A straight line drawn from the transmitting antenna to the visible horizon of this larger earth will indicate the true radio line-of-sight distance. The difference in the behavior of light and radio waves in this frequency range is due almost entirely to the difference in wavelength.

#### Effects of Reflections of Radio Waves

The degree of reflection will depend on the conductivity of the obstacle and its physical size as compared with the wavelength. Reflections from objects of good conductivity are particularly strong, and since many of the works of men are

metallic, reflections from them often are troublesome in television transmission. Whenever the waves of the desired signal reach the receiving antenna over paths of different lengths, there is the possibility of multipath interference.

Obviously, when the difference in path length is great enough, the receiver receives two signal impulses carrying the same information, but separated by a short time interval. If the two signals are of approximately the same amplitude, the interference thus produced can be serious. When television signals from multipath reception are received, a ghost image, or images, is produced. This image is displaced to the right of the main image on the face of the picture tube.

#### Common Types of Television Antennas

**DIPOLE WITH REFLECTOR.** Although the simple half-wave antenna will provide satisfactory reception within reasonable distances from the transmitter, in outlying areas the signals become weaker and are affected more by noise and interference. In order to increase the signal gain of the antenna, a reflector is added to the antenna. Not only is the gain of the antenna increased, but the width of the response area is decreased (Figure 15-21). Therefore, a small displacement of the antenna one way or the other alters the strength of the received signal appreciably. This property is useful in preventing reflected signals from reaching the receiver.

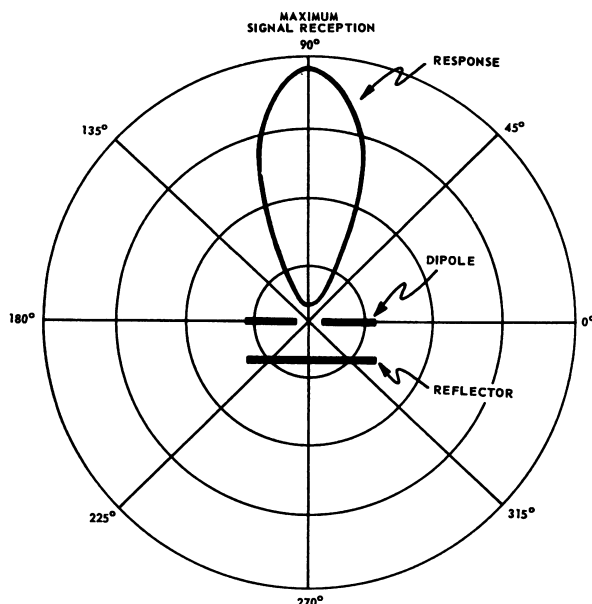


Figure 15-21. Directional Response Curve of a Dipole Antenna with Reflector

**FOLDED DIPOLE.** Another type of antenna in wide use is the folded-dipole antenna. The folded

dipole has the same bidirectional pattern as the simple dipole and approximately the same gain, but the impedance is close to 300 ohms as compared with the 73 ohms of the simple dipole. The response of the folded dipole, however, is more uniform over a band of frequencies than the simple dipole. The addition of a reflector will increase the directivity of the folded dipole.

**CONICAL ANTENNA.** The conical antenna shown in Figure 15-22 is one of a large number of special antennas that have been developed to operate satisfactorily over a wide frequency band. The front elements are veered forward while the rear elements (the reflectors) generally extend straight out. Due to the position of the elements, the response pattern of the antenna contains a major lobe on each channel. One array is satisfactory for all VHF channels.

**STACKED ANTENNAS.** In some areas where the signal strength is low, the gain of the receiving antenna can be increased by *stacking* the basic dipoles, as shown in Figure 15-23. This antenna is sometimes known as the "Lazy H" because it looks like the letter H lying on its side. The two dipoles are mounted one above the other; the

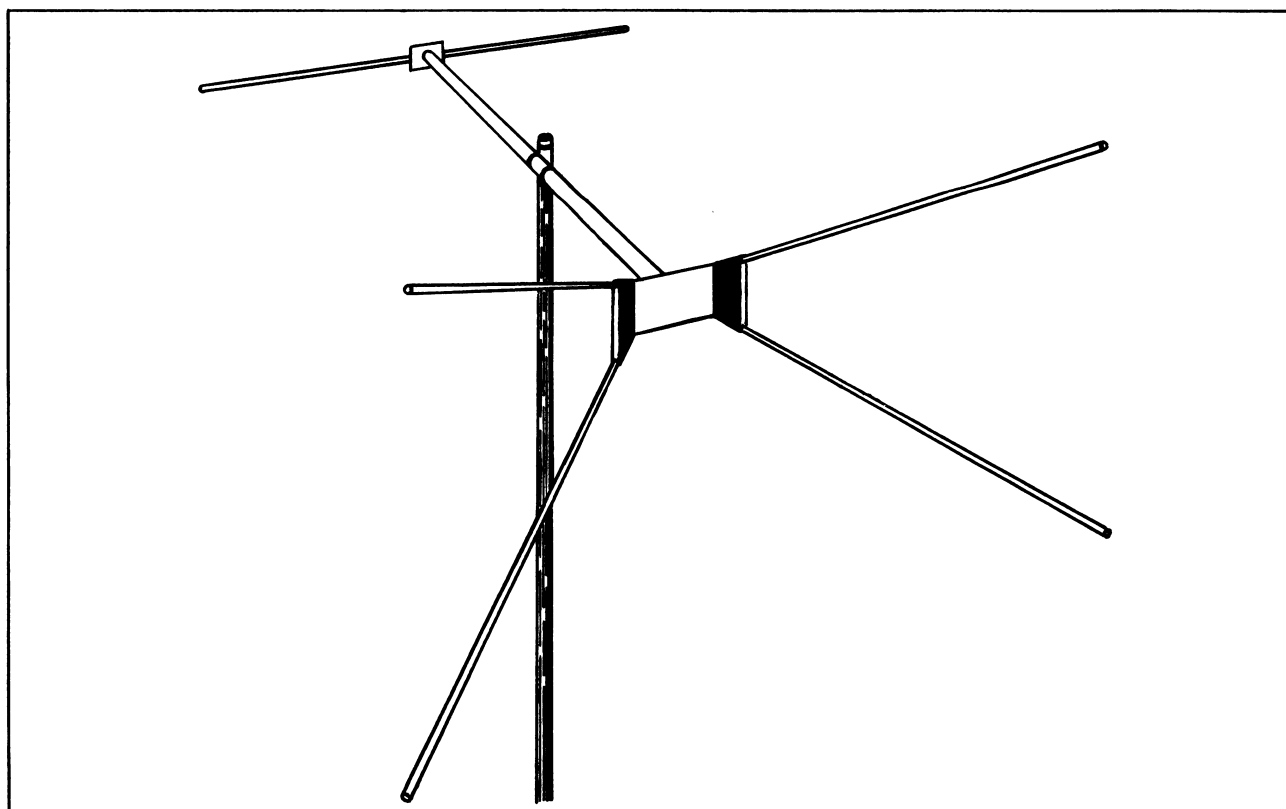


Figure 15-22. Conical Antenna



center terminals of each dipole are connected with a parallel wire. Each conductor of the transmission line is attached to this parallel wire at a point midway between the dipoles. A reflector can be placed behind each dipole to make the array more directive. Also, any one of the basic antennas can be stacked to provide greater gain.

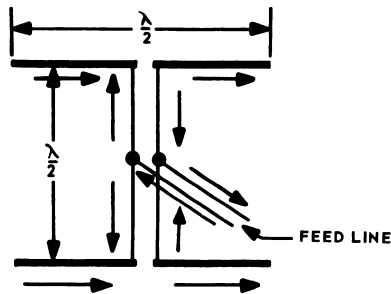


Figure 15-23. Stacked Dipole Antenna

**MULTIELEMENT ANTENNAS.** The directional response of an antenna can be further improved by adding directors to the array (Figure 15-24). Whereas the reflector is slightly longer than the dipole and placed behind it, the directors are slightly shorter than the dipole and placed in front of it. This arrangement is known as the Yagi antenna; it has a sharp unidirectional response pattern. The radiator may be either a

simple or a folded dipole, and the number of directors may vary from one to twelve or more.

### Transmission Lines

A transmission line is used to couple energy from the transmitter to the transmitting antenna, and from the receiving antenna to the receiver. Its purpose is to transfer power from one to the other with least possible loss. How well this purpose is accomplished depends on the characteristics of the transmission line used. For maximum transfer of power the characteristic impedance of the transmission line should match the characteristic impedance of the antenna.

Normally, it is most convenient and economical to use an antenna that is capable of receiving all the VHF television stations. This requires a fairly uniform response from 54 to 216 mc. Since a resonant dipole has an impedance of 73 ohms at its center, the connecting transmission line should match this value to effect a maximum transfer of power. However, a dipole cut to resonate at 50 mc will have an impedance of 2000 ohms at 100 mc, hence a transmission line of higher impedance is required.

Since line loss is inversely proportional to the characteristic impedance, it is desirable to use as high an impedance value as possible. However, the size of the line must be considered; a compromise has been effected so that in current

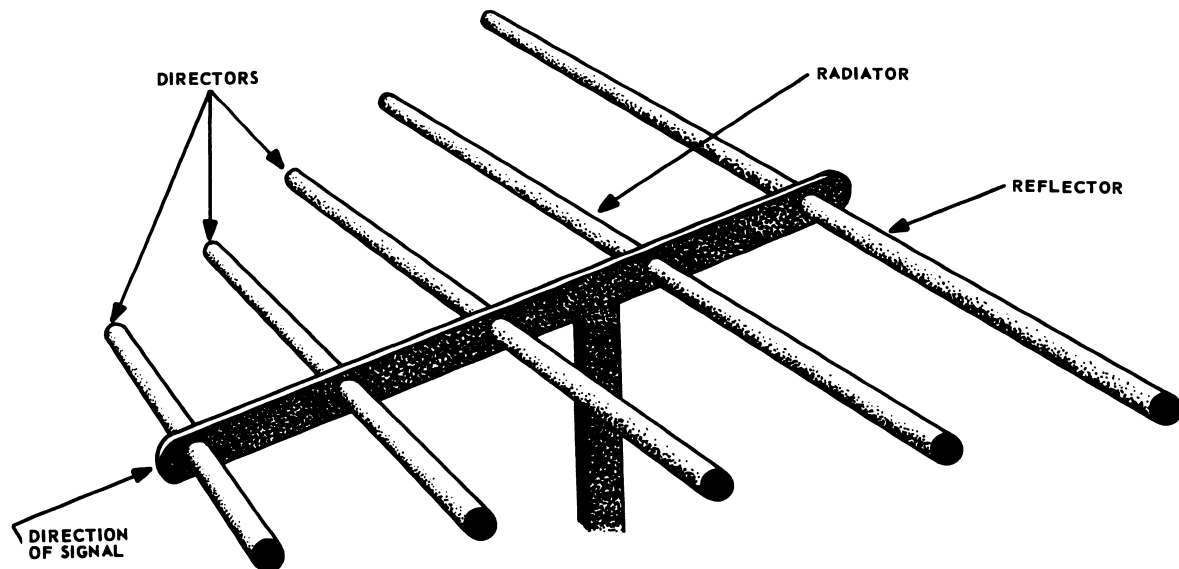


Figure 15-24. Yagi Antenna

practice a standard 300-ohm transmission line is used. It has been found that a 300-ohm line used with a dipole produces a broad frequency response without excessive loss due to mismatching. Transmission lines that are used in television were included in Figures 10-21, 10-22, and 10-24, Chapter 10.

After the correct antenna and transmission line have been chosen, the installation must be made with certain principles in mind. The higher the antenna, the stronger the signal received. Before fixing the supports permanently, the antenna should be tested by connecting it to the receiver. When more than one station is to be received, the final placement of the antenna must be a compromise. In extreme cases, it may be necessary to install more than one antenna.

### UHF Antennas

The addition of the UHF band to the original VHF band created new problems in television communications. At the higher frequencies, transmission line losses are larger and noise becomes a greater problem in the receiver. The first UHF receivers fed the signals directly to a crystal mixer, where they were converted to a lower frequency and then amplified. Since the early UHF receivers did not use r-f preamplifiers before the mixer, they were normally less sensitive than VHF receivers. However, new tube design has made it possible to use preamplifiers in UHF receivers. This increases the sensitivity.

Because of the poorer receiver sensitivity, UHF antennas must be erected more carefully than VHF antennas. Not only is it necessary to select a good location, but the proper height must also be determined. However, high-gain antennas are more feasible at UHF than at VHF because the antenna dimensions are smaller. Thus, a dipole at 800 mc will have approximately one-tenth the over-all dimensions of a dipole at 80 mc. This means that more elements can be added to the UHF array without causing it to be unwieldy. Therefore, since the gain of an antenna generally rises with the number of elements, higher gain can be anticipated from UHF arrays.

### Types of UHF Antennas

**FAN DIPOLE.** As in the VHF band, the simplest type of antenna in the UHF band is the dipole.

In order to provide the broad band characteristics necessary to receive all signals within the television UHF band, triangular sheets of metal are used instead of rods, as shown in Figure 15-25A. The array is normally designed with a corner angle of  $70^\circ$  in order to have an input impedance of approximately 300 ohms. The radiation pattern of a fan dipole is a figure-eight, but if a screen is placed behind the dipole as shown in Figure 15-25B, the field becomes unidirectional. For greatest gain, the over-all length of the fan dipole should be about 16 inches.

Since the gain is only slightly greater than that of a rod dipole, these units provide satisfactory reception in strong signal areas where there are relatively few ghost signals. However, fan dipoles can be stacked two and four high to provide increased gain and better discrimination against ground-reflected signals. Notice that the reflector

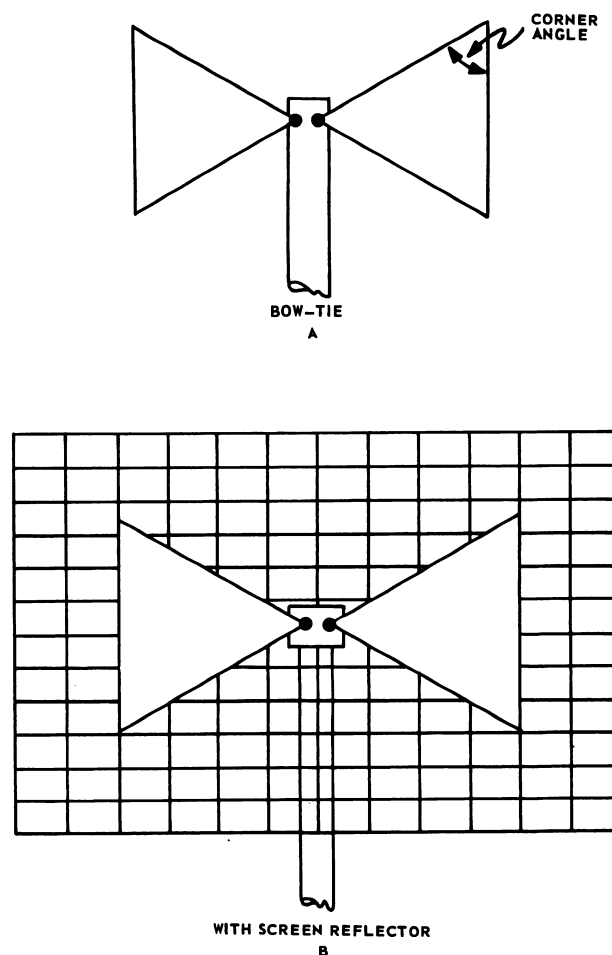


Figure 15-25. Fan Dipole Antenna

in Figure 15-25B is a wire screen instead of the rods usually used at VHF. Screens are much more efficient reflectors than rods, but are seldom used at VHF because they are too bulky at the lower frequencies. Screens are as effective as solid metallic reflectors provided the openings are no larger than two-tenths of a wavelength at the highest operating frequency. Reflector dimensions are not critical, but the edges should extend for a short distance beyond the dipole elements.

**RHOMBIC ANTENNAS.** The rhombic antenna shown in Figure 15-26 has been used successfully for VHF and UHF reception in weak signal areas. It has good gain, broad bandpass characteristics, and a sharply defined unidirectional response pattern. Since an increase in the length of the sides, or legs, results in higher gain and a sharper directivity pattern, the length of each side should be at least several wavelengths.

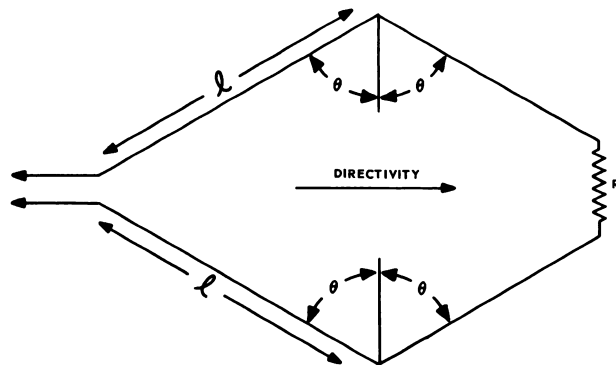


Figure 15-26. Rhombic Antenna

In order to obtain the unidirectional effect, a resistor must be inserted at one end of the array. For use with UHF television receivers, the resistor will have a value of 470 to 500 ohms. The impedance reflected into the transmission line is slightly less than the terminating resistance and will match a 300-ohm line without a serious loss of gain. In weak signal areas, rhombic antennas can be stacked for greater gain. The array is then oriented with the resistor end pointing in the direction of the station.

**CYLINDRICAL PARABOLIC REFLECTOR.** A parabolic reflector can be used to obtain a high concentration of electromagnetic energy. This was explained in Chapter 13. The use of the reflector

increases the gain and directivity. It is possible to use only a section of a parabolic reflector. Adding a cylindrical parabolic reflector to a resonant dipole results in a gain of about 8 db. This array will provide good results where high gain is desired and the ghost problem is not serious.

**CORNER REFLECTORS.** Two flat surfaces can also be used to form a reflector as shown in Figure 15-27. This type of antenna is called the *corner reflector* antenna. The receiving element is usually a dipole placed at the center of the corner angle but at some distance from the vertex. The response pattern of this array not only depends

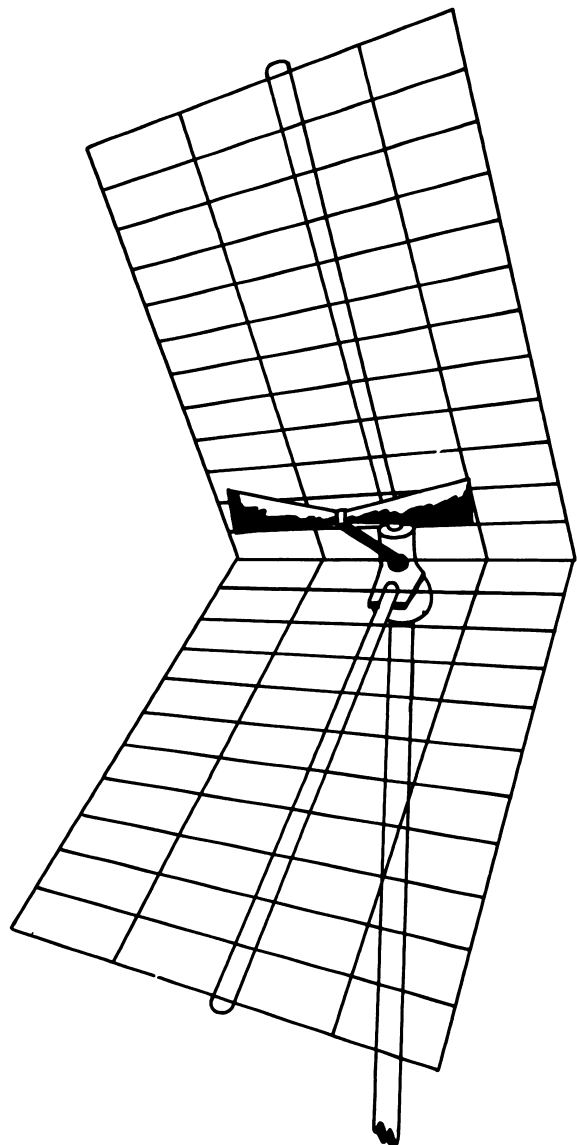


Figure 15-27. Fan Dipole with Corner Reflector

on the corner angle but also on the distance between the dipole and the reflector corner. If the dipole is located too far from the vertex, the response pattern will have several lobes. If it is brought in too close, the vertical response will be broadened, and the array will become more susceptible to ground-reflected signals.

**OTHER ANTENNAS.** Actually, any of the VHF antennas can be used for UHF television if the size is reduced to the proper dimensions. The design formulas remain the same and no other changes are required. However, because of the smaller dimensions, stacked arrays are more commonly found among UHF antennas. This is understandable since greater gain is obtainable from the increased number of elements.

### COLOR TELEVISION

Engineers have been experimenting with color television since 1946. Several systems were submitted, but were rejected because they were unsuitable for commercial use. The line and field frequencies were different from those of the black and white receivers. The National Television Systems Committee (NTSC) system was finally adopted in 1953.

#### Color Elements

If one is to see the true color of an object, it must be seen under a light which has all of the wavelengths of light. The color of an object is the function of the wavelengths of light that are not absorbed by the object. Anyone who has worked in a theater with projector lamps has found that when different colored lights are combined, the resulting color, as seen by the audience, is different in hue from any of the projected

lights. Experimentation has found that the colors green, red, and blue, when combined in various proportions, will produce a wide range of colors. These three colors work better, in this respect, than any other combination of three colors, and have been chosen as the primary colors of television.

#### The National Television Systems Committee (NTSC) System

This system was designed so that the signal would not occupy more than 6 mc, carrying both the entire black and white signal and the color information. It was found that in the 4 megacycles assigned to the video signal, video information was not contained in every cycle. Instead, the monochrome signal appeared in clusters of energy that were located close to the harmonics of the 15,750-cycle line-scanning frequency. This meant that empty spaces were available for additional information. This is illustrated in Figure 15-28.

To make the second signal fall between the clusters of energy of the first signal and thus cause no interference, odd multiples of 7,875 (half of 15,750) are used. This placing of clusters of energy of one signal between clusters of energy of another signal is called interleaving (Figure 15-29). From this we see that the entire color signal is actually composed of two components — the monochrome signal and the color information signal.

The monochrome signal (black and white portion of the entire signal), for all practical purposes, is the same as the black and white signal in non-color sets. This signal is produced by the combination of the green, blue, and red signals in certain proportions.

$$Y = 0.60G + 0.30R + 0.10B$$

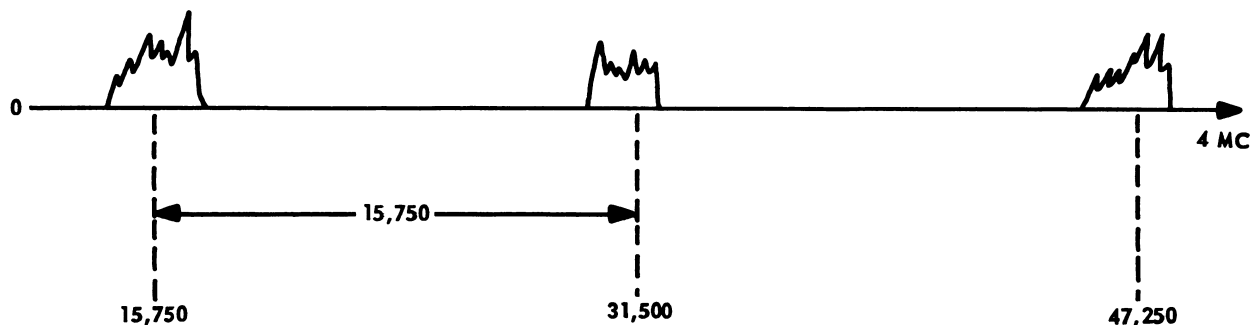


Figure 15-28. Spectrum Analysis of Monochrome Signal

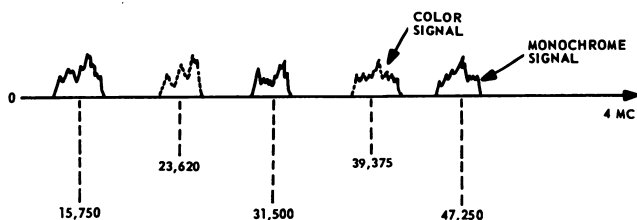


Figure 15-29. Interleaving of Signals

The letter *Y* is used to denote the monochrome signal; the letters *G*, *R* and *B* stand for green, red, and blue. To the human eye, green is brighter than red and much brighter than blue. That is why these proportions were chosen. Only the variations in amplitude of the picture signal are contained in the monochrome video signal, and it is these variations that give us the light intensity changes at the picture tube screen.

The color information signal is interleaved with the monochrome signal. We determine the amount of information this component must carry to form the color image from the reaction of the eye to color. The average human eye, when viewing a relatively large area, sees a full color range; but when the area decreases, distinguishing between colors becomes difficult. When the area is very small, the eye can only detect changes in brightness. The NTSC system utilizes these color seeing characteristics of the eye by sending the large and medium size areas in color and the very small areas in black and white. We find, too, that more color is given to the large objects and areas than to the medium size areas.

The first step in transmitting color is to produce a type of subcarrier, with a frequency of approximately 3.58 mc, and its associated sidebands. The next step is to modulate this subcarrier with the proper voltages representing the three colors — red, green, and blue.

In the color signal, we find that the *Y* signal (monochrome) has components from 0 to 4 mc and that the subcarrier (color) frequency is 3.58 mc. Two color signals modulate the subcarrier; these are designated as the *I* and *Q* signals. The *I* color signal (red) frequencies extend from 0 to 1.5 mc; the *Q* color signal (blue) frequencies extend from 0 to 0.5 mc.

It is sufficient to transmit two color signals, *I* and *Q*, as the *Y* signal contains voltages from all three colors.

The color subcarrier is suppressed in the NTSC system to reduce the interference that would be formed in connection with the sound carrier, a 920-kc beat frequency. When the image is televised in black and white, the entire color signal is removed.

A disadvantage of a suppressed subcarrier is that it must be reinserted in the receiver so that detection can be accomplished. It is important that the subcarrier have the correct phase. The information concerning the frequency and phase of the subcarrier is provided in a color burst that is transmitted with the signal. This burst is located on the "back porch" of each blanking pulse. The information in this burst is used to synchronize the frequency and phase of the 3.58-mc oscillator in the receiver.

### Receivers

Circuits in the color receiver which are relatively the same as those in the black and white receiver are r-f tuner, mixer, video i-f, sound i-f, audio detector, audio amplifiers, video detector, video amplifiers, sync circuits, and high-voltage power supply. Those circuits which are not common are chrominance circuits, color sync circuits, and the tri-gun color picture tube and associated circuits.

There are a relatively large number of stages in the chrominance channel. These stages are used to demodulate the color signal and reobtain the original *I* and *Q* voltages. The *I* and *Q* signals, with the *Y* signal, are mixed in a matrix; the resulting voltages represent the red, green, and blue components in the *I*, *Q*, and *Y* signals. After amplification, the three voltages are fed to the respective control grids of the tri-gun color tube.

The color sync section has the distinct function of generating a 3.58-mc signal that possesses the proper phase relationship to the transmitted subcarrier. The necessary information is obtained from the color burst that is located on the "back porch" of each horizontal sync pulse. If the phase relationship is wrong, there is a very noticeable form of distortion.

The tri-gun color picture tube uses three electron guns, while the black and white picture tube uses only one electron gun. The screen of the black and white tube has only one type of phosphor, while the tri-gun color tube has three different color-emitting phosphors. While globules were used on the mosaic image plate of the camera

tube, phosphor dots are used on the tri-gun color tube. They are placed in an orderly arrangement of triangular groups to form approximately 340,000 trios of dots. Each of the electron guns activates only one type of phosphor. We see the light that is given off by the phosphor when it is activated by the electron beam; the color of light we see is determined by the phosphors that are being struck by the electron beam, and by the number of electrons in each beam.

Since any electron beam could activate any phosphor dot on the screen, there must be some way of allowing each beam to strike only one type of phosphor dot. This is accomplished by using a shadow mask. The mask is in front of, and parallel to, the screen; it contains circular holes equal in number to the dot trios. The holes and electron guns are so aligned, with respect to each group, that an electron gun beam can strike only one phosphor dot of a trio. The other two dots are concealed by the shadow mask. Each beam strikes one phosphor dot of the trio. The associated convergence circuits insure that each beam is properly converged at every point of the screen.

The components that are mounted externally on the tri-gun color picture tube are a little different from those mounted on the black-and-white tube. The deflection yoke is more complex because three beams are being deflected instead of one. Another component is the purity coil or magnet. This component adjusts the axis of each electron beam so that the beam will strike the correct phosphor dot. Proper alignment of each beam with respect to phosphor dot plate and shadow mask is the function of the purity coil. The convergence coils, as stated previously, insure that each beam is properly converged at every point on the screen. Ion traps are not used in this tube because the color screen is aluminized. The layer of aluminum is the barrier to the onrushing ions.

## CLOSED CIRCUIT TELEVISION

### Components of Video Signal Generating Equipment

Basically, the closed circuit system consists of the following equipments: pickup cameras, camera control equipment, monitors, synchronizing generator, projectors and multiplexers, and a parallel audio system.

Since motion picture film is frequently used in closed circuit TV, television film projectors are an integral part of a television system. They must be able to operate continuously for long periods, start and stop almost instantly without serious speed variations, and run at constant speed for long periods of time.

Television projectors produce a sharp, stable, optical image from the film and project it with sufficient intensity of light on the photosensitive surface of the camera pickup tube. Either 16-mm or 35-mm motion picture film or slides can be used. Black and white or color, positive or negative, film may be used.

Motion picture film has a standard projection rate of 24 frames per second, while television systems operate at 30 frames per second. Therefore, film projectors for television must be designed to synchronize the 24-frame film with the 30-frame television system. This will permit any standard 16-mm or 35-mm film to be shown. In addition, the television film projector must be synchronized with the 60-cycle power line if it is used with pickup cameras such as the iconoscope.

Multiplexers make possible the use of several film and slide projectors with only one pickup camera. The multiplexer is the optical link that selects one of several film, slide, or opaque projector outputs for projection into a film pickup camera. Movable mirrors, driven by motors or relays, change the optical paths of the several optical picture sources so that the selected picture may be directed to the camera pickup tube. The mirrors, motors, and control panel are all in one unit, or housing, which constitutes the multiplexer.

### Video Signal Distribution Equipment

After a television signal has been produced and processed, it must be distributed or transmitted to the video signal display equipment. This is accomplished by the video signal distribution equipment. It includes distribution amplifiers, equalizers, r-f generators, r-f distribution, microwave relay, and switching equipments.

Distribution amplifiers are divided into two groups: (1) video distribution and (2) pulse distribution amplifiers. Their function is to drive the television signals through the low-impedance coaxial cables. They must pass the signal with



the least distortion, provide gain (or amplification), insure isolation between input and output, and add the sync signal to the video signal to form the composite television signal.

Video equalizers are inserted into the coaxial lines to correct for distortion introduced by the cables. Coaxial cables have relatively poor amplitude-versus-frequency response. Long cables or improperly terminated cables may cause high-frequency attenuation and accompanying phase distortion. Such distortion will cause reduced resolution, ringing, smearing, or streaks. Equalizers are composed of variable combinations of inductors, capacitors, and resistors. They are designed to complement the cable characteristics and to produce an amplitude-versus-frequency response that is uniform. Although correction of poor amplitude response is the primary use, video equalizers will correct most of the phase distortion also.

Coaxial cable distributes the television signal to the immediate area; for remote distribution, however, r-f generators and r-f distribution systems are required. R-f generators are low-power transmitters usually tuned to one of the VHF channels. Instead of feeding an antenna, they feed a coaxial line that carries the VHF signal to one or more television receivers. The video signals modulate the output of the r-f generator. The first type of r-f generator accepts only video signals and uses a single television channel. The second type will accept both video and associated audio signals, and can usually operate on any of the standard VHF television channels. These r-f generators have the capability of accepting both video and audio signals simultaneously and distributing them on a single coaxial cable to one or more television receivers.

The r-f distribution system is a highly versatile and effective means of distributing a number of video signals to several standard receivers. The number of receivers used is limited to the length of coaxial line used and the degree of isolation desired between receivers.

Microwave equipment is used for the distribution of television signals over distances of 5 miles or more, where the use of coaxial cable is difficult and costly. A television microwave relay system actually is a complete television system, containing transmitter, transmitting antenna, receiving antenna, and receiver. The system is designed to operate at UHF frequencies ranging

from 1000 mc to 15,000 mc. With a minimum of transmitter power and highly directional antennas, it is possible to relay television signals over distances of 30 miles or more without serious attenuation or distortion.

Typical applications of a microwave relay system are from studio to transmitter, from studio to a remote building, between two widely separated buildings, on a large Air Force installation, and from mobile studio to main studio.

Switching equipment is a necessity in any but the simplest, one-camera television system. The equipment selects the output from one of several signal sources and connects it to one outgoing path. In a complex system the switching equipment provides for program monitoring and previewing; video switching between studio cameras, film cameras and remote pickups; and special effects such as fading and lap dissolving.

Video switchers use pushbutton switching and may operate either directly or remotely by relay.

A zero time interval during transfer is ideal, but as this is not practical, the switching equipment is designed to make the transfer by using either gap switching or overlap. Both of these methods permit picture transfer with a minimum of disturbance. Also, there are several other functions that the switching equipment performs. By the use of synchronization, the transfer is smoothly accomplished with fade-ins and fade-outs, lap dissolves, and diagonal wipes. The equipment also provides for two signals to be superimposed, or for two signals to be mixed in split screen montages. Provisions are also provided for previewing the video and for monitoring the output line.

The following components are included in the switching equipment: basic switching unit (relay chassis), fader assembly, gain and isolation stages, special effects and lap dissolve amplifiers (may be fed back to input for previewing), and mixing or output amplifiers (synchronization is added).

Television monitors take a video frequency signal and reproduce a picture image. The signal may be taken directly from a video line or from a demodulated signal from a television receiver. By the addition of r-f and i-f stages, the monitor may function as a receiver. The size of the monitor picture tube is determined by its particular function. A 21-inch tube may be used in a classroom or office; a 17-inch tube may be used in a

studio control room; an 8-inch tube is sufficient for cueing work or continuity checks.

### Kinescope Recorders

Kinescope recording is the name given to the recording of television programs on motion picture film. This is often called video tape. Special equipment was designed in order to have recordings of high quality for closed-circuit and other applications. A kinescope recording system normally includes a high quality television monitor, a specially designed motion picture camera, and the necessary equipment for recording the sound simultaneously with the picture.

A requirement of the kinescope recorder is that it must synchronize with the television system and resolve the timing difference between it and the TV system. Other requirements are adequate light from the monitor kinescope, and camera exposure time accurately controlled and synchronized with the television system. The motion picture cameras, either 16-mm or 35-mm, should

be capable of producing either negative or positive recorded film prints. The complete equipment should have a high degree of reliability and should be ruggedly built for continuous operation.

### Typical Closed-Circuit Video System

A closed-circuit system may consist of simply one camera and one monitor combination or it may be quite extensive, including several studios feeding program material to many different locations for viewing. A one-camera, one-monitor system is limited in its usage for different programming, but it might be adequate for a function such as simple surveillance or telemetering.

A moderate closed-circuit system, which is comparable to a small television broadcasting station, is shown in Figure 15-30. Different camera angles and subjects are available for alternate showing without any breaks in program continuity. This system is suitable to Air Force application in classrooms and briefing rooms, and where there are no requirements for film repro-

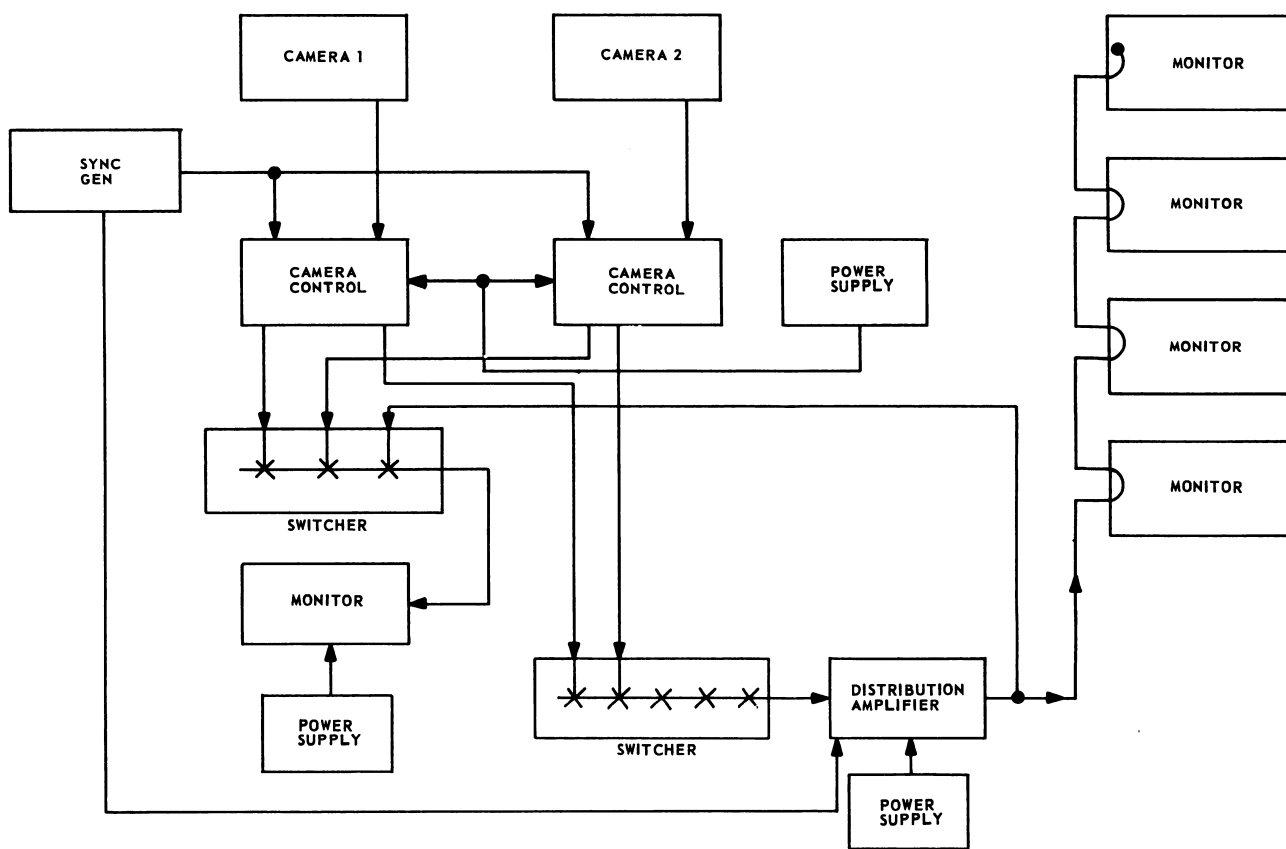


Figure 15-30. Moderate Closed-Circuit Television System

duction. A moderate variety of program material can be presented through this system.

A complete closed-circuit system contains the necessary equipment to provide a versatile, top-quality program. The complete system is comparable to a medium-sized commercial broadcasting station, or to an elaborate closed-circuit system used in some colleges and universities. The complete system is acquired by adding a camera, film reproduction equipment, a video tape recorder, and the necessary distribution equipment to the moderate system. Like the moderate system, the complete system is excellent for classroom application and has an increased capability for a very efficient presentation of a technical lecture with illustrations.

A multiple studio system is the same as the two previously described with the exception of added equipment. The main addition is that of one or more studios, with the distribution, control, and switching equipment that is a part of a studio.

### **Audio System**

As previously stated, the audio system is completely separate from the video equipment in a television system. Film projectors, video tape recorders, and television receivers are the only pieces of equipment in the video system that have any audio stages or functions. In the planning of a television system, this separation of audio and video often leads to a neglect of the audio. The video installation will be limited in scope if the sound facilities are not adequate. In the planning of a closed-circuit system, future expansion should be anticipated as well as the satisfying of present needs for audio and video facilities.

The audio requirements parallel the different video systems. Just as the simple television system consists of one pickup camera and one monitor, a simple audio system consists of one microphone and one speaker. For the presentation of varied program material, as technical instruction, a more complex audio system is demanded. The system should include facilities for intercommunication, program audio, and talkback.

### **Applications of Closed-Circuit Television**

There are two major classes of application of closed circuit TV: program television and telemetering, or remote monitoring. The dividing

line between the two is not always completely distinct.

Under program television, there are broadcasting, educational, kinescope recording, and communications. Under telemetering and remote monitoring, there are hazardous and inaccessible locations, data remoting and collection, magnification, ultraviolet, X-ray, and infrared sources. We will discuss here the types the Air Force is concerned with.

Educational TV has an enormous potential. The consensus of leading educators is that a carefully prepared course taught by television has many advantages over conventional classroom techniques. This assumes that all possible equipment and potential of television programming is used.

Kinescope recording has many advantages over direct photography. Through the use of kinescope recording and television techniques, it is possible to produce certain types of films more cheaply than by direct photography.

Community system is the use of a community antenna system where receivers in a town are connected by cable to a central distribution box. This system could well be used in closed-circuit for programs produced locally.

Closed-circuit communications may include several types of applications. These include the use of television at large meetings and intercity hookups.

Examples of hazardous or inaccessible locations where TV might be used are inside of an aircraft wing, the observation of missile firing, in atomic radiation areas, and in high temperatures and high noise level areas.

In data remoting and collection, information is transmitted from a point where it would be expensive or inconvenient to obtain it directly. Information can be brought in from several points to a single viewing location and, conversely, it is possible to bring information from a single control location to a number of branch points. Examples are — use in military operations control centers, and weather briefings.

Image magnification can be used in classroom explanations on small parts or adjustments in electronic equipment. Long focal length lenses make it possible for small objects to be brought closer for more detailed observation. Thus, small objects can be viewed by a large number of students.

TV makes possible the observation of objects illuminated by X-ray, ultraviolet, and infrared sources.

### **Specific Applications**

Only a small cross section of the tasks that closed-circuit television can perform have been given. Some areas in which tasks can be performed by closed-circuit TV are: (1) training and instruction, (2) security, (3) general surveillance, (4) traffic control, (5) weather service, and (6) hazardous monitoring. Each area will be discussed briefly with specific applications in each area.

**TRAINING AND INSTRUCTION.** The Air Force has found closed-circuit TV to be of much value in its military schools. A greater number of students can be served by the best instructor available by using closed-circuit TV. Television is an excellent visual aid as it combines the use of charts, slides, and films. The value of television as a teaching aid is unquestioned. For the present, television is accepted as a supplement to, rather than a replacement for, conventional classroom instruction.

**SECURITY.** Continued surveillance of a restricted area is normally done by uniformed guards. Sometimes it is desirable to conceal the surveillance. A hidden closed-circuit system can do this task much better and more consistently than the best continuous surveillance of guards. Other areas where security can be checked by a hidden camera are offices, file rooms, mail rooms, entrances and exits.

**GENERAL SURVEILLANCE.** While closed-circuit TV is used in this area principally by industry at the present time, military applications may be expected to include those closely paralleling the civilian applications. Examples of general surveillance are data remoting, radar remoting, administrative use, medical applications, hospital wards, Air Police applications, aerial reconnaissance, ground reconnaissance, warehouse func-

tions, fire warning surveillance, and personnel supervision.

**TRAFFIC CONTROL.** A very good usage of closed-circuit television for the Air Force is that of observing, recording, and directing of aircraft ground movement. Railroad switching and vehicular movement are also expedited by television. Air traffic control towers rely upon visual observation. Television equipment would extend the vision of the personnel in several ways. All parts of the field would be visible.

**WEATHER SERVICE.** Flight crew briefings on weather station data is greatly facilitated by the use of closed-circuit television. The forecaster has all his data always at hand and only the briefest preparation is necessary. The forecasting map and the winds aloft chart are the principal information sources that demand picture transmitting equipment.

### **Color Closed-Circuit Television**

As in monochrome closed-circuit television, color closed-circuit television has unlimited capabilities. Since the addition of color closely approximates reality, the introduction of color into television as an educational medium presents to its users the most versatile type of closed-circuit educational program possible. In technical training programs, the use of color is a necessity rather than a luxury. Especially is this true in the fields of electronics, communications, weather, aircraft engine repair, and hydraulics.

Color plays a major role in providing identification for small parts, subassemblies, test points, and control functions in each of these technical areas. Without color, oral instructions, via television, would require an extensive amount of explanation. As the need for newer and faster methods of technical training is ever increasing, the introduction of color closed-circuit television could change this condition and could create a new standard in military technical instruction.

BY ORDER OF THE SECRETARY OF THE AIR FORCE

OFFICIAL

R. J. PUGH  
*Colonel, USAF*  
*Director of Administrative Services*

CURTIS E. LEMAY  
*Chief of Staff*

# Symbols and Formulas

## ELECTROSTATICS

$$F = \frac{Q_1 Q_2}{K d^2} \quad \text{Force between charges}$$

$$E = \frac{Q}{K d^2} \quad \text{Electric field intensity}$$

$$E = \frac{V}{d} \quad \text{Field between plates}$$

$$F = QE \quad \text{Force on } Q \text{ in field } E$$

$$W = QV = eV \quad \text{Work}$$

$$C = \frac{Q}{V} \quad \text{Capacitance}$$

$$V = \frac{Q}{d} \quad \text{Potential}$$

## DIRECT CURRENT

$$E = IR \quad \text{Ohm's Law}$$

$$P = EI = I^2 R = \frac{E^2}{R} \quad \text{Power}$$

### Series Circuits

$$R_t = R_1 + R_2 + R_3 + \dots \quad \text{Total resistance in a series circuit}$$

$$E_t = E_1 + E_2 + E_3 + \dots \quad \text{Total voltage in a series circuit}$$

$$I_t = I_1 = I_2 = I_3 = \dots \quad \text{Total current in a series circuit}$$

$$P_t = P_1 + P_2 + P_3 + \dots \quad \text{Total power in a series circuit}$$

### Parallel Circuits

$$R_t = \frac{R_1 R_2}{R_1 + R_2} \quad \text{Total resistance — Two resistors in parallel}$$

$$R_t = \frac{R_1 R_2}{R_1 + R_2} \quad \text{Unknown resistance — Two resistors in parallel with total resistance known}$$

$$G_t = \frac{1}{R_t} \quad \text{Total conductance}$$

$$G_t = G_1 + G_2 \quad \text{Total conductance}$$

$$R_t = \frac{1}{\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} + \dots} \quad \text{Total resistance with any number of resistors in parallel}$$

$$E = E_1 = E_2 = E_3 = \dots \quad \text{Total voltage in a parallel circuit}$$

$$I_t = I_1 + I_2 + I_3 + \dots \quad \text{Total current in a parallel circuit}$$

$$P_t = P_1 + P_2 + P_3 + \dots \quad \text{Total power in a parallel circuit}$$

### Resistance

$$R = P \frac{l}{d^2} \quad \text{Value of resistance as a function of the material and its physical dimensions}$$

where  $R$  = resistance in ohms  
 $P$  = resistance in ohms per circular-mil-foot of the material (specific resistance)  
 $l$  = length of conductor in feet  
 $d$  = diameter of wire in mils

### Meters

$$R_{\text{shunt}} = \frac{I_m R_m}{I_t - I_m} \quad \text{Shunt required to extend range of ammeter}$$

where  $I_m$  = current for full scale deflection  
 $R_m$  = meter resistance  
 $I_t$  = extended range of current

$$\text{Ohms per volt} = \frac{\text{Resistance of meter}}{\text{Full scale reading in volts}} \times \text{Sensitivity of a meter}$$

$$\text{or} = \frac{1}{\text{full scale current}}$$

$$R_m = (E \times (E \times \Omega/V) - R_v) \quad \text{Multiplier required to extend range of voltmeter}$$

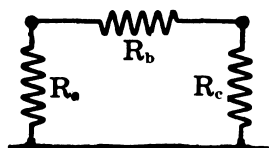
where  $E$  = extended range of voltage

$\Omega/V$  = sensitivity of meter

$R_v$  = meter resistance

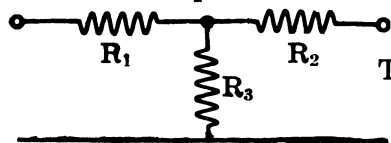
### Networks

$$R_1 = \frac{R_a R_b}{R_a + R_b + R_c}$$



$\pi$  to T transformation

$$R_a = \frac{R_1 R_2 + R_2 R_3 + R_3 R_1}{R_2}$$



T to  $\pi$  transformation

### MAGNETISM AND ELECTROMAGNETS

$$F = \frac{m_1 m_2}{\mu d^2} \quad \text{Force between two magnetic poles}$$

where  $\mu$  = permeability of the medium

$$H = \frac{F}{m} \quad \text{Magnetic field intensity}$$

$$H = \frac{2I}{10r} \quad \text{Magnetic intensity about a conductor}$$

where  $I$  = current in amperes  
 $r$  = radius in centimeters

$$H = \frac{4\pi NI}{10 l} \quad \text{Magnetic intensity for long coils}$$

$$H = \frac{2\pi NI}{10r} \quad \text{Magnetic intensity for short coils}$$

$$\text{mmf} = \frac{4\pi NI}{10} \quad \text{gilberts} \quad \text{Magnetomotive force}$$

$$\phi = \frac{\text{mmf}}{R} \quad \text{maxwells} \quad \text{Flux}$$

$$R = \frac{L}{\mu A} \quad \text{rels} \quad \text{Reluctance}$$

$$B = \frac{\phi}{A} \quad \text{gauss} \quad \text{Flux Density}$$

$$F = \frac{B I l}{10} \sin \theta \quad \text{dynes} \quad \text{Force exerted on conductor}$$

$$\mu = \frac{\beta}{H} \quad \text{Permeability}$$

### Induction

$$e = -N \frac{\Delta \phi}{\Delta t} \times 10^{-8} \quad \text{Average Induced voltage}$$

$$e = -M \frac{\Delta i}{\Delta t} \quad \text{Induced voltage}$$

where  $M$  is mutual inductance

$$M = K \sqrt{L_1 L_2} \quad \text{Mutual Inductance}$$

where  $k$  is the coefficient of coupling

$$L = \frac{N\phi}{I} \times 10^{-8} \quad \text{Inductance}$$

where  $N$  is number of turns  
 $I$  is current in amperes

$$e_L = -L \frac{\Delta i}{\Delta t} \quad \text{Voltage across inductance}$$

$$e_L = -N \frac{\Delta \phi}{\Delta t} \times 10^{-8} \quad \text{Voltage across inductance}$$

### ALTERNATING CURRENT

#### General

$$e = E_{\max} \sin \omega t \quad \text{Instantaneous voltage}$$

where  $\omega = 2\pi f$



$$i = I_{\max} \sin \omega t \quad \text{Instantaneous current}$$

$$\text{where } \omega = 2\pi f$$

$$P = \frac{E^2}{R} = I^2 R = EI \cos \theta \quad \text{Power in AC circuits}$$

where  $\theta$  is the phase angle

$$PF = \frac{R}{Z} = \cos \theta \quad \text{Power factor}$$

$$\theta = \arctan \frac{X}{R} = \text{angle of lead or lag (degrees)} \quad \text{Phase angle}$$

$$C = K \frac{A}{d} \quad \text{Capacitance in micromicrofarads}$$

where A is area of plates

d is distance between plates

K is dielectric constant

$$X_L = 2\pi f L \quad \text{Inductive reactance}$$

$$X_C = \frac{1}{2\pi f C} \quad \text{Capacitive reactance}$$

$$Z = \sqrt{R^2 + (X_L - X_C)^2} \quad \text{Impedance — triangular form}$$

$$Z = R \pm jX \quad \text{Impedance — rectangular form}$$

$$F_r = \frac{1}{2\pi \sqrt{LC}} \quad \text{Frequency at resonance}$$

$$Q = \frac{X_L}{R} \quad \text{Merit of a coil or tank}$$

where R is total resistance in series with coil

$$Q = \frac{R}{X_L} \quad \text{Merit of a coil or tank}$$

where R is total resistance in parallel with coil

$$L_T = L_1 + L_2 \pm 2M \quad \text{Total inductance of two coils}$$

where M is the mutual inductance

$$L_t = L_1 + L_2 + L_3 \dots \quad \text{Inductance in series}$$

$$1/L_t = 1/L_1 + 1/L_2 + 1/L_3 \dots \quad \text{Inductance in parallel}$$

$$C_t = C_1 + C_2 + C_3 \dots \quad \text{Capacitances in parallel}$$

$$1/C_t = 1/C_1 + 1/C_2 + 1/C_3 \dots \quad \text{Capacitances in series}$$

$$Q_t = Q_1 = Q_2 = Q_3 \dots \quad \text{Total charge on condensers in series}$$

$$Q_t = Q_1 + Q_2 + Q_3 \dots \quad \text{Total charge on condensers in parallel}$$

+ is used when the fields add  
- is used when the fields oppose

### Transformers

$$\frac{E_p}{E_s} = \frac{N_p}{N_s} = \frac{I_s}{I_p} \quad \text{Relationships among voltages, turns, and currents in transformers}$$

$$\frac{Z_p}{Z_s} = \left( \frac{N_p}{N_s} \right)^2 \quad \text{Impedances and turns}$$

$$Z = Z_1 + Z_p - \frac{Z_m^2}{Z_2 + Z_s} \quad \text{Impedance when looking into primary of a transformer}$$

where  $Z_1$  is in series with primary  
 $Z_p$  is impedance of primary windings

$Z_m$  is mutual impedance

$Z_s$  is impedance of secondary windings

$Z_2$  is impedance in series with secondary

$$Z_r = \frac{-Z_m^2}{Z_2 + Z_s} \quad \text{Reflected impedance}$$

### Transients

$$t = R \times C \quad \text{Time constant for RC circuits}$$

where t is in seconds

R is in ohms

C is in farads

$$t = \frac{L}{R} \quad \text{Time constant for L/R circuits}$$

where t is in seconds

R is in ohms

L is in henries

$$e_c = E_a \left( 1 - e^{-\frac{t}{RC}} \right) \quad \text{Voltage change across a capacitor in an RC circuit}$$

where  $E_a$  is the applied voltage

$$e_R = E_a e^{-\frac{t}{RC}} \quad \text{Voltage across a resistor in an RC circuit}$$

$$i = \frac{E_a}{R} e^{-\frac{t}{RC}} \quad \text{Current in an RC circuit}$$

$$e_L = E_a e^{-\frac{Rt}{L}} \quad \text{Voltage across coil in an L/R circuit}$$

$$e_R = E_a \left( 1 - e^{-\frac{Rt}{L}} \right) \quad \text{Voltage change across a resistor in an L/R circuit}$$

$$i = \frac{E_a}{R} \left( 1 - e^{-\frac{Rt}{L}} \right) \quad \text{Current in an L/R circuit}$$

$$E_R = (E_a - E_{co}) e^{-\frac{t}{RC}} \quad \text{Voltages in a direct current restorer}$$

where  $E_{co}$  is the voltage on the capacitor at the start

#### VACUUM TUBES AND AMPLIFIERS

$$\mu = \frac{\Delta e_p}{\Delta e_g} \quad (i_p \text{ constant}) \quad \text{Amplification factor}$$

$$r_p = \frac{\Delta e_p}{\Delta i_p} \quad (e_g \text{ constant}) \quad \text{Plate resistance}$$

where  $r_p$  is in ohms  
 $e_p$  is in volts  
 $i_p$  is in amperes

$$g_m = \frac{\Delta i_p}{\Delta e_g} \quad (e_p \text{ constant}) \quad \text{Transconductance}$$

where  $g_m$  is in mhos  
 $i_p$  is in amperes  
 $e_g$  is in volts

$$\mu = g_m r_p \quad \text{Relationship among } g_m, r_p \text{ and } \mu$$

where  $g_m$  is in mhos  
 $r_p$  is in ohms

$$A = \frac{E_o}{E_s} = \frac{E_o}{E_g} = \frac{\mu Z_L}{r_p + Z_L} \quad \text{Voltage gain of triode operated Class A}$$

where  $E_o$  is the output voltage

$E_s$  is the input voltage

$E_g$  is the grid-to-cathode voltage

$Z_L$  is the equivalent load impedance

$r_p$  is the plate resistance

$$db = 20 \log \frac{E_o}{E_{in}} = 20 \log \frac{I_o}{I_{in}} \quad \text{Voltage or current gain in decibels, if } R_{in} \text{ equals } R_{out}$$

$$db = 10 \log \frac{P_o}{P_{in}} \quad \text{Power gain of stage operated Class A}$$

where  $P_o$  = output power

$P_{in}$  = input power

$$A_t = A_1 \times A_2 \times A_3 \dots A_n \quad \text{Total voltage gain of stages in cascade}$$

$$db_t = db_1 + db_2 + db_3 \dots db_n \quad \text{Total power gain of stages in cascade}$$

$$F_h = \frac{1}{2\pi R_L C_s} \quad \text{Highest frequency in band passed by a video amplifier}$$

where  $R_L$  is the total load resistance  
 $C_s$  is total shunt capacitance

$$F_l = \frac{1}{2\pi R_g C_c} \quad \text{Lowest frequency in band passed by a video amplifier}$$

where  $R_g$  is the grid resistor  
 $C_c$  is the coupling capacitor

$$A' = \frac{E_o}{E_g} = \frac{A}{1 - BA} \quad \text{Gain in a feedback amplifier}$$

where  $A$  is the gain without feedback  
 $B$  is the fraction of the output voltage that is fed back

#### Cathode Followers

$$Z_o = \frac{R_k r_p}{r_p + R_k (\mu + 1)} \quad \text{Output impedance of a cathode follower}$$

$$A = \frac{E_o}{E_s} = \frac{\mu R_k}{r_p + R_k (\mu + 1)} \quad \text{Gain of a cathode follower}$$

## OSCILLATORS

$$F_o = \frac{1}{2\pi\sqrt{LC}} \quad \text{Output frequency of LC oscillator}$$

$$F_o = \frac{1}{2\pi\sqrt{R_1 C_1 R_2 C_2}} \quad \text{Output frequency of a Wien bridge oscillator}$$

$$F_o = \frac{1}{2\pi R_1 C_1} \quad \text{Output frequency of a Wien bridge oscillator when } R_1 C_1 = R_2 C_2$$

$$F_o = \frac{1}{2\pi RC\sqrt{6}} \quad \text{Output frequency of a phase shift oscillator}$$

### Shock Excited Oscillators

$$Q = \frac{R_s}{X_L} = \frac{R_s}{L/C} \quad \text{Merit of oscillator tank}$$

where  $R_s$  is the shunt resistance

$$X_L = X_C = \sqrt{L/C} \text{ at resonance}$$

$$F_o = \frac{1}{2\pi\sqrt{LC}} \quad \text{Frequency of oscillator when } Q \text{ is 5 or greater}$$

$$F_o = \frac{1}{2\pi\sqrt{LC}} \sqrt{1 - \left(\frac{1}{2Q}\right)^2} \quad \text{Frequency of oscillator when } Q \text{ is less than 5}$$

$$E_{\max} = I_p \sqrt{L/C} \quad \text{Maximum output voltage}$$

## TRANSMISSION LINES

$$Z_o = \sqrt{L/C} = \frac{E_i}{I_i} \quad \text{Characteristic impedance}$$

where  $E_i$  is the incident voltage  
 $I_i$  is the incident current

$$SWR = \frac{E_{\max}}{E_{\min}} = \frac{I_{\max}}{I_{\min}} = \sqrt{\frac{Z_{\max}}{Z_{\min}}} = \frac{1 + |r|}{1 - |r|} \quad \text{Standing wave ratio}$$

$$\text{where } |r| = \frac{p - 1}{p + 1}$$

$$SWR = \frac{Z_L}{Z_o} \quad \text{Standing wave ratio (Resistive load only)}$$

where  $Z_L$  is larger than  $Z_o$

$$SWR = \frac{Z_o}{Z_L} \quad \text{Standing wave ratio (Resistive load only)}$$

where  $Z_o$  is larger than  $Z_L$

$$Z_o = \sqrt{Z_{in} Z_L} \quad \text{Impedance of matching transformer}$$

where  $Z_{in}$  is input impedance  
 $Z_o$  = characteristic impedance  
 $Z_L$  is output impedance

$$t = n \sqrt{LC} \quad \text{Time for a wave to travel one length of a delay line}$$

where  $n$  is the number of sections in the line

$$PW = 2n \sqrt{LC} \quad \text{Pulse width of a pulse forming line}$$

$$f_{co} = \frac{1}{\pi \sqrt{LC}} \quad \text{Cutoff frequency of an artificial line}$$

## WAVE GUIDES AND CAVITY RESONATORS

$$\lambda_a = k \frac{300}{f_{mc}} \quad \text{Wavelength in free space (air)}$$

where  $k$  is the velocity correction factor  
 $f_{mc}$  is frequency in megacycles

$$V_p = f\lambda_g \quad \text{Phase velocity}$$

where  $\lambda_g$  is wavelength in guide

$$V_a = f\lambda_a = \sqrt{V_p V_g} \quad \text{Velocity in free space}$$

where  $\lambda_a$  is wavelength in free space (air)

$$\lambda_c = \frac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2}} \quad \text{Wavelength at cutoff}$$

where  $m$  is the first subscript in TE or TM mode designations

$n$  is the second subscript in TE or TM mode designations

$a$  is the short dimension of the wave guide

$b$  is the long dimension of the wave guide

## MICROWAVE OSCILLATORS

$$F_e = 1.6 \times 10^{-12} E$$

Force in dynes acting  
on an electron in  
an electric field

where E is the electric field  
strength in volts/cm

$$F_m = 1.6 \times 10^{-20} V_p B$$

Force in dynes acting  
on an electron in  
a magnetic field

where  $V_p$  is the component of  
electron velocity perpendicular  
to the direction of the magnetic  
field

B is magnetic flux density in  
gausses

$$K.E. = \frac{1}{2} mv^2 = eV$$

Kinetic energy in ergs

where m is mass of electron in  
grams

v is velocity of electron  
in cm/sec

e is charge on electron  
in coulomb

V is potential difference  
through which the electron  
travels

# Tables

## Powers of Ten

|            |   |                |
|------------|---|----------------|
| $10^{10}$  | = | 10,000,000,000 |
| $10^9$     | = | 1,000,000,000  |
| $10^8$     | = | 100,000,000    |
| $10^7$     | = | 10,000,000     |
| $10^6$     | = | 1,000,000      |
| $10^5$     | = | 100,000        |
| $10^4$     | = | 10,000         |
| $10^3$     | = | 1,000          |
| $10^2$     | = | 100            |
| $10^1$     | = | 10             |
| $10^0$     | = | 1              |
| $10^{-1}$  | = | .1             |
| $10^{-2}$  | = | .01            |
| $10^{-3}$  | = | .001           |
| $10^{-4}$  | = | .000 1         |
| $10^{-5}$  | = | .000 01        |
| $10^{-6}$  | = | .000 001       |
| $10^{-7}$  | = | .000 000 1     |
| $10^{-8}$  | = | .000 000 01    |
| $10^{-9}$  | = | .000 000 001   |
| $10^{-10}$ | = | .000 000 000 1 |

# COMMON LOGARITHMS

| N  | 0    | 1    | 2    | 3    | 4    | 5    | 6    | 7    | 8    | 9    |
|----|------|------|------|------|------|------|------|------|------|------|
| 0  | .... | 0000 | 3010 | 4771 | 6021 | 6990 | 7782 | 8451 | 9031 | 9542 |
| 1  | 0000 | 0414 | 0792 | 1139 | 1461 | 1761 | 2041 | 2304 | 2553 | 2788 |
| 2  | 3010 | 3222 | 3424 | 3617 | 3802 | 3979 | 4150 | 4314 | 4472 | 4624 |
| 3  | 4771 | 4914 | 5051 | 5185 | 5315 | 5441 | 5563 | 5682 | 5798 | 5911 |
| 4  | 6021 | 6128 | 6232 | 6335 | 6435 | 6532 | 6628 | 6721 | 6812 | 6902 |
| 5  | 6990 | 7076 | 7160 | 7243 | 7324 | 7404 | 7482 | 7559 | 7634 | 7709 |
| 6  | 7782 | 7853 | 7924 | 7993 | 8062 | 8129 | 8195 | 8261 | 8325 | 8388 |
| 7  | 8451 | 8513 | 8573 | 8633 | 8692 | 8751 | 8808 | 8865 | 8921 | 8976 |
| 8  | 9031 | 9085 | 9138 | 9191 | 9243 | 9294 | 9345 | 9395 | 9445 | 9494 |
| 9  | 9542 | 9590 | 9638 | 9685 | 9731 | 9777 | 9823 | 9868 | 9912 | 9956 |
| 10 | 0000 | 0043 | 0086 | 0128 | 0170 | 0212 | 0253 | 0294 | 0334 | 0374 |
| 11 | 0414 | 0453 | 0492 | 0531 | 0569 | 0607 | 0645 | 0682 | 0719 | 0755 |
| 12 | 0792 | 0828 | 0864 | 0899 | 0934 | 0969 | 1004 | 1038 | 1072 | 1106 |
| 13 | 1139 | 1173 | 1206 | 1239 | 1271 | 1303 | 1335 | 1367 | 1399 | 1430 |
| 14 | 1461 | 1492 | 1523 | 1553 | 1584 | 1614 | 1644 | 1673 | 1703 | 1732 |
| 15 | 1761 | 1790 | 1818 | 1847 | 1875 | 1903 | 1931 | 1959 | 1987 | 2014 |
| 16 | 2041 | 2068 | 2095 | 2122 | 2148 | 2175 | 2201 | 2227 | 2253 | 2279 |
| 17 | 2304 | 2330 | 2355 | 2380 | 2405 | 2430 | 2455 | 2480 | 2504 | 2529 |
| 18 | 2553 | 2577 | 2601 | 2625 | 2648 | 2672 | 2695 | 2718 | 2742 | 2765 |
| 19 | 2788 | 2810 | 2833 | 2856 | 2878 | 2900 | 2923 | 2945 | 2967 | 2989 |
| 20 | 3010 | 3032 | 3054 | 3075 | 3096 | 3118 | 3139 | 3160 | 3181 | 3201 |
| 21 | 3222 | 3243 | 3263 | 3284 | 3304 | 3324 | 3345 | 3365 | 3385 | 3404 |
| 22 | 3424 | 3444 | 3464 | 3483 | 3502 | 3522 | 3541 | 3560 | 3579 | 3598 |
| 23 | 3617 | 3636 | 3655 | 3674 | 3692 | 3711 | 3729 | 3747 | 3766 | 3784 |
| 24 | 3802 | 3820 | 3838 | 3856 | 3874 | 3892 | 3909 | 3927 | 3945 | 3962 |
| 25 | 3979 | 3997 | 4014 | 4031 | 4048 | 4065 | 4082 | 4099 | 4115 | 4133 |
| 26 | 4150 | 4166 | 4183 | 4200 | 4216 | 4232 | 4249 | 4265 | 4281 | 4298 |
| 27 | 4314 | 4330 | 4346 | 4362 | 4378 | 4393 | 4409 | 4425 | 4440 | 4456 |
| 28 | 4472 | 4487 | 4502 | 4518 | 4533 | 4548 | 4564 | 4579 | 4594 | 4609 |
| 29 | 4624 | 4639 | 4654 | 4669 | 4683 | 4698 | 4713 | 4728 | 4742 | 4757 |
| 30 | 4771 | 4786 | 4800 | 4814 | 4829 | 4843 | 4857 | 4871 | 4886 | 4900 |
| 31 | 4914 | 4928 | 4942 | 4955 | 4969 | 4983 | 4997 | 5011 | 5024 | 5038 |
| 32 | 5051 | 5065 | 5079 | 5092 | 5105 | 5119 | 5132 | 5145 | 5159 | 5172 |
| 33 | 5185 | 5198 | 5211 | 5224 | 5237 | 5250 | 5263 | 5276 | 5289 | 5302 |
| 34 | 5315 | 5328 | 5340 | 5353 | 5366 | 5378 | 5391 | 5403 | 5416 | 5428 |
| 35 | 5441 | 5453 | 5465 | 5478 | 5490 | 5502 | 5514 | 5527 | 5539 | 5551 |
| 36 | 5563 | 5575 | 5587 | 5599 | 5611 | 5623 | 5635 | 5647 | 5658 | 5670 |
| 37 | 5682 | 5694 | 5705 | 5717 | 5729 | 5740 | 5752 | 5763 | 5775 | 5786 |
| 38 | 5798 | 5809 | 5821 | 5832 | 5843 | 5855 | 5866 | 5877 | 5888 | 5899 |
| 39 | 5911 | 5922 | 5933 | 5944 | 5955 | 5966 | 5977 | 5988 | 5999 | 6010 |
| 40 | 6021 | 6031 | 6042 | 6053 | 6064 | 6075 | 6085 | 6096 | 6107 | 6117 |
| 41 | 6128 | 6138 | 6149 | 6160 | 6170 | 6180 | 6191 | 6201 | 6212 | 6222 |
| 42 | 6232 | 6243 | 6253 | 6263 | 6274 | 6284 | 6294 | 6304 | 6314 | 6325 |
| 43 | 6335 | 6345 | 6355 | 6365 | 6375 | 6385 | 6395 | 6405 | 6415 | 6425 |
| 44 | 6435 | 6444 | 6454 | 6464 | 6474 | 6484 | 6493 | 6503 | 6513 | 6522 |
| 45 | 6532 | 6542 | 6551 | 6561 | 6571 | 6580 | 6590 | 6599 | 6609 | 6618 |
| 46 | 6628 | 6637 | 6646 | 6656 | 6665 | 6675 | 6684 | 6693 | 6702 | 6712 |
| 47 | 6721 | 6730 | 6739 | 6749 | 6758 | 6767 | 6776 | 6785 | 6794 | 6803 |
| 48 | 6812 | 6821 | 6830 | 6839 | 6848 | 6857 | 6866 | 6875 | 6884 | 6893 |
| 49 | 6902 | 6911 | 6920 | 6928 | 6937 | 6946 | 6955 | 6964 | 6972 | 6981 |
| 50 | 6990 | 6998 | 7007 | 7016 | 7024 | 7033 | 7042 | 7050 | 7059 | 7067 |
| N  | 0    | 1    | 2    | 3    | 4    | 5    | 6    | 7    | 8    | 9    |



## COMMON LOGARITHMS

| N   | 0    | 1    | 2    | 3    | 4    | 5    | 6    | 7    | 8    | 9    |
|-----|------|------|------|------|------|------|------|------|------|------|
| 50  | 6990 | 6998 | 7007 | 7016 | 7024 | 7033 | 7042 | 7050 | 7059 | 7067 |
| 51  | 7076 | 7084 | 7093 | 7101 | 7110 | 7118 | 7126 | 7135 | 7143 | 7152 |
| 52  | 7160 | 7168 | 7177 | 7185 | 7193 | 7202 | 7210 | 7218 | 7226 | 7235 |
| 53  | 7243 | 7251 | 7259 | 7267 | 7275 | 7284 | 7292 | 7300 | 7308 | 7316 |
| 54  | 7324 | 7332 | 7340 | 7348 | 7356 | 7364 | 7372 | 7380 | 7388 | 7396 |
| 55  | 7404 | 7412 | 7419 | 7427 | 7435 | 7443 | 7451 | 7459 | 7466 | 7474 |
| 56  | 7482 | 7490 | 7497 | 7505 | 7513 | 7520 | 7528 | 7536 | 7543 | 7551 |
| 57  | 7559 | 7566 | 7574 | 7582 | 7589 | 7597 | 7604 | 7612 | 7619 | 7627 |
| 58  | 7634 | 7642 | 7649 | 7657 | 7664 | 7672 | 7679 | 7686 | 7694 | 7701 |
| 59  | 7709 | 7716 | 7723 | 7731 | 7738 | 7745 | 7752 | 7760 | 7767 | 7774 |
| 60  | 7782 | 7789 | 7796 | 7803 | 7810 | 7818 | 7825 | 7832 | 7839 | 7846 |
| 61  | 7853 | 7860 | 7868 | 7875 | 7882 | 7889 | 7896 | 7903 | 7910 | 7917 |
| 62  | 7924 | 7931 | 7938 | 7945 | 7952 | 7959 | 7966 | 7973 | 7980 | 7987 |
| 63  | 7993 | 8000 | 8007 | 8014 | 8021 | 8028 | 8035 | 8041 | 8048 | 8055 |
| 64  | 8062 | 8069 | 8075 | 8082 | 8089 | 8096 | 8102 | 8109 | 8116 | 8122 |
| 65  | 8129 | 8136 | 8142 | 8149 | 8156 | 8162 | 8169 | 8176 | 8182 | 8189 |
| 66  | 8195 | 8202 | 8209 | 8215 | 8222 | 8228 | 8235 | 8241 | 8248 | 8254 |
| 67  | 8261 | 8267 | 8274 | 8280 | 8287 | 8293 | 8299 | 8306 | 8312 | 8319 |
| 68  | 8325 | 8331 | 8338 | 8344 | 8351 | 8357 | 8363 | 8370 | 8376 | 8382 |
| 69  | 8388 | 8395 | 8401 | 8407 | 8414 | 8420 | 8426 | 8432 | 8439 | 8445 |
| 70  | 8451 | 8457 | 8463 | 8470 | 8476 | 8482 | 8488 | 8494 | 8500 | 8506 |
| 71  | 8513 | 8519 | 8525 | 8531 | 8537 | 8543 | 8549 | 8555 | 8561 | 8567 |
| 72  | 8573 | 8579 | 8585 | 8591 | 8597 | 8603 | 8609 | 8615 | 8621 | 8627 |
| 73  | 8633 | 8639 | 8645 | 8651 | 8657 | 8663 | 8669 | 8675 | 8681 | 8686 |
| 74  | 8692 | 8698 | 8704 | 8710 | 8716 | 8722 | 8727 | 8733 | 8739 | 8745 |
| 75  | 8751 | 8756 | 8762 | 8768 | 8774 | 8779 | 8785 | 8791 | 8797 | 8802 |
| 76  | 8808 | 8814 | 8820 | 8825 | 8831 | 8837 | 8842 | 8848 | 8854 | 8859 |
| 77  | 8865 | 8871 | 8876 | 8882 | 8887 | 8893 | 8899 | 8904 | 8910 | 8915 |
| 78  | 8921 | 8927 | 8932 | 8938 | 8943 | 8949 | 8954 | 8960 | 8965 | 8971 |
| 79  | 8976 | 8982 | 8987 | 8993 | 8998 | 9004 | 9009 | 9015 | 9020 | 9025 |
| 80  | 9031 | 9036 | 9042 | 9047 | 9053 | 9058 | 9063 | 9069 | 9074 | 9079 |
| 81  | 9085 | 9090 | 9096 | 9101 | 9106 | 9112 | 9117 | 9122 | 9128 | 9133 |
| 82  | 9138 | 9143 | 9149 | 9154 | 9159 | 9165 | 9170 | 9175 | 9180 | 9186 |
| 83  | 9191 | 9196 | 9201 | 9206 | 9212 | 9217 | 9222 | 9227 | 9232 | 9238 |
| 84  | 9243 | 9248 | 9253 | 9258 | 9263 | 9269 | 9274 | 9279 | 9284 | 9289 |
| 85  | 9294 | 9299 | 9304 | 9309 | 9315 | 9320 | 9325 | 9330 | 9335 | 9340 |
| 86  | 9345 | 9350 | 9355 | 9360 | 9365 | 9370 | 9375 | 9380 | 9385 | 9390 |
| 87  | 9395 | 9400 | 9405 | 9410 | 9415 | 9420 | 9425 | 9430 | 9435 | 9440 |
| 88  | 9445 | 9450 | 9455 | 9460 | 9465 | 9469 | 9474 | 9479 | 9484 | 9489 |
| 89  | 9494 | 9499 | 9504 | 9509 | 9513 | 9518 | 9523 | 9528 | 9533 | 9538 |
| 90  | 9542 | 9547 | 9552 | 9557 | 9562 | 9566 | 9571 | 9576 | 9581 | 9586 |
| 91  | 9590 | 9595 | 9600 | 9605 | 9609 | 9614 | 9619 | 9624 | 9628 | 9633 |
| 92  | 9638 | 9643 | 9647 | 9652 | 9657 | 9661 | 9666 | 9671 | 9675 | 9680 |
| 93  | 9685 | 9689 | 9694 | 9699 | 9703 | 9708 | 9713 | 9717 | 9722 | 9727 |
| 94  | 9731 | 9736 | 9741 | 9745 | 9750 | 9754 | 9759 | 9763 | 9768 | 9773 |
| 95  | 9777 | 9782 | 9786 | 9791 | 9795 | 9800 | 9805 | 9809 | 9814 | 9818 |
| 96  | 9823 | 9827 | 9832 | 9836 | 9841 | 9845 | 9850 | 9854 | 9859 | 9863 |
| 97  | 9868 | 9872 | 9877 | 9881 | 9886 | 9890 | 9894 | 9899 | 9903 | 9908 |
| 98  | 9912 | 9917 | 9921 | 9926 | 9930 | 9934 | 9939 | 9943 | 9948 | 9952 |
| 99  | 9956 | 9961 | 9965 | 9969 | 9974 | 9978 | 9983 | 9987 | 9991 | 9996 |
| 100 | 0000 | 0004 | 0009 | 0013 | 0017 | 0022 | 0026 | 0030 | 0035 | 0039 |
| N   | 0    | 1    | 2    | 3    | 4    | 5    | 6    | 7    | 8    | 9    |

# NATURAL LOGARITHMS

| N   | 0.0    | 1.0    | 2.0    | 3.0    | 4.0    | 5.0    | 6.0    | 7.0    | 8.0    | 9.0    |
|-----|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
|     | ....   | 0.0000 | 0.6931 | 1.0986 | 1.3863 | 1.6094 | 1.7918 | 1.9459 | 2.0794 | 2.1972 |
| 10  | 2.3026 | 2.3979 | 2.4849 | 2.5649 | 2.6391 | 2.7081 | 2.7726 | 2.8332 | 2.8904 | 2.9444 |
| 20  | 9957   | 3.0445 | 3.0910 | 3.1355 | 3.1781 | 3.2189 | 3.2581 | 3.2958 | 3.3322 | 3.3673 |
| 30  | 3.4012 | 4340   | 4657   | 4965   | 5264   | 5553   | 5835   | 6109   | 6376   | 6636   |
| 40  | 6889   | 7136   | 7377   | 7612   | 7842   | 8067   | 8286   | 8501   | 8712   | 8918   |
| 50  | 9120   | 9318   | 9512   | 9703   | 9890   | 4.0073 | 4.0254 | 4.0431 | 4.0604 | 4.0775 |
| 60  | 4.0943 | 4.1109 | 4.1271 | 4.1431 | 4.1589 | 1744   | 1897   | 2047   | 2195   | 2341   |
| 70  | 2485   | 2627   | 2767   | 2905   | 3041   | 3175   | 3307   | 3408   | 3567   | 3694   |
| 80  | 3820   | 3944   | 4067   | 4188   | 4308   | 4427   | 4543   | 4659   | 4773   | 4886   |
| 90  | 4998   | 5109   | 5218   | 5326   | 5433   | 5539   | 5643   | 5747   | 5850   | 5951   |
| 100 | 6052   | 6151   | 6250   | 6347   | 6444   | 6540   | 6634   | 6728   | 6821   | 6913   |
| 110 | 7005   | 7095   | 7185   | 7274   | 7362   | 7449   | 7536   | 7622   | 7707   | 7791   |
| 120 | 7875   | 7958   | 8040   | 8122   | 8203   | 8283   | 8363   | 8442   | 8520   | 8598   |
| 130 | 8675   | 8752   | 8828   | 8903   | 8978   | 9053   | 9127   | 9200   | 9273   | 9345   |
| 140 | 9416   | 9488   | 9558   | 9628   | 9698   | 9767   | 9836   | 9904   | 9972   | 5.0039 |
| 150 | 5.0106 | 5.0173 | 5.0239 | 5.0304 | 5.0370 | 5.0434 | 5.0499 | 5.0562 | 5.0626 | 0689   |
| 160 | 0752   | 0814   | 0876   | 0938   | 0999   | 1059   | 1120   | 1180   | 1240   | 1299   |
| 170 | 1358   | 1417   | 1475   | 1533   | 1591   | 1648   | 1705   | 1761   | 1818   | 1874   |
| 180 | 1930   | 1985   | 2040   | 2095   | 2149   | 2204   | 2257   | 2311   | 2364   | 2417   |
| 190 | 2470   | 2523   | 2575   | 2627   | 2679   | 2730   | 2781   | 2832   | 2883   | 2933   |
| 200 | 2983   | 3033   | 3083   | 3132   | 3181   | 3230   | 3279   | 3327   | 3375   | 3423   |
| 210 | 3471   | 3519   | 3566   | 3613   | 3660   | 3706   | 3753   | 3799   | 3845   | 3891   |
| 220 | 3936   | 3982   | 4027   | 4072   | 4116   | 4161   | 4205   | 4250   | 4293   | 4337   |
| 230 | 4381   | 4424   | 4467   | 4510   | 4553   | 4596   | 4638   | 4681   | 4723   | 4765   |
| 240 | 4806   | 4848   | 4889   | 4931   | 4972   | 5013   | 5053   | 5094   | 5134   | 5175   |
| 250 | 5215   | 5255   | 5294   | 5334   | 5373   | 5413   | 5452   | 5491   | 5530   | 5568   |
| 260 | 5607   | 5645   | 5683   | 5722   | 5759   | 5797   | 5835   | 5872   | 5910   | 5947   |
| 270 | 5984   | 6021   | 6058   | 6095   | 6131   | 6168   | 6204   | 6240   | 6276   | 6312   |
| 280 | 6348   | 6384   | 6419   | 6454   | 6490   | 6525   | 6560   | 6595   | 6630   | 6664   |
| 290 | 6699   | 6733   | 6768   | 6802   | 6836   | 6870   | 6904   | 6937   | 6971   | 7004   |
| 300 | 7038   | 7071   | 7104   | 7137   | 7170   | 7203   | 7236   | 7268   | 7301   | 7333   |
| 310 | 7366   | 7398   | 7430   | 7462   | 7494   | 7526   | 7557   | 7589   | 7621   | 7652   |
| 320 | 7683   | 7714   | 7746   | 7777   | 7807   | 7838   | 7869   | 7900   | 7930   | 7961   |
| 330 | 7991   | 8021   | 8051   | 8081   | 8111   | 8141   | 8171   | 8201   | 8230   | 8260   |
| 340 | 8289   | 8319   | 8348   | 8377   | 8406   | 8435   | 8464   | 8493   | 8522   | 8551   |
| 350 | 8579   | 8608   | 8636   | 8665   | 8693   | 8721   | 8749   | 8777   | 8805   | 8833   |
| 360 | 8861   | 8889   | 8916   | 8944   | 8972   | 8999   | 9026   | 9054   | 9081   | 9108   |
| 370 | 9135   | 9162   | 9189   | 9216   | 9243   | 9269   | 9296   | 9322   | 9349   | 9375   |
| 380 | 9402   | 9428   | 9454   | 9480   | 9506   | 9532   | 9558   | 9584   | 9610   | 9636   |
| 390 | 9661   | 9687   | 9713   | 9738   | 9764   | 9789   | 9814   | 9839   | 9865   | 9890   |
| 400 | 9915   | 9940   | 9965   | 9989   | 6.0014 | 6.0039 | 6.0064 | 6.0088 | 6.0113 | 6.0137 |
| 410 | 6.0162 | 6.0186 | 6.0210 | 6.0234 | 0259   | 0283   | 0307   | 0331   | 0355   | 0379   |
| 420 | 0403   | 0426   | 0450   | 0474   | 0497   | 0521   | 0544   | 0568   | 0591   | 0615   |
| 430 | 0638   | 0661   | 0684   | 0707   | 0730   | 0753   | 0776   | 0799   | 0822   | 0845   |
| 440 | 0868   | 0890   | 0913   | 0936   | 0958   | 0981   | 1003   | 1026   | 1048   | 1070   |
| 450 | 1092   | 1115   | 1137   | 1159   | 1181   | 1203   | 1225   | 1247   | 1269   | 1291   |
| 460 | 1312   | 1334   | 1356   | 1377   | 1399   | 1420   | 1442   | 1463   | 1485   | 1506   |
| 470 | 1527   | 1549   | 1570   | 1591   | 1612   | 1633   | 1654   | 1675   | 1696   | 1717   |
| 480 | 1738   | 1759   | 1779   | 1800   | 1821   | 1841   | 1862   | 1883   | 1903   | 1924   |
| 490 | 1944   | 1964   | 1985   | 2005   | 2025   | 2046   | 2066   | 2086   | 2106   | 2126   |
| 500 | 2146   | 2166   | 2186   | 2206   | 2226   | 2246   | 2265   | 2285   | 2305   | 2324   |
| N   | 0.0    | 1.0    | 2.0    | 3.0    | 4.0    | 5.0    | 6.0    | 7.0    | 8.0    | 9.0    |

**NATURAL LOGARITHMS**

| <b>N</b>    | <b>0.0</b> | <b>1.0</b> | <b>2.0</b> | <b>3.0</b> | <b>4.0</b> | <b>5.0</b> | <b>6.0</b> | <b>7.0</b> | <b>8.0</b> | <b>9.0</b> |
|-------------|------------|------------|------------|------------|------------|------------|------------|------------|------------|------------|
| <b>500</b>  | 6.2146     | 6.2166     | 6.2186     | 6.2206     | 6.2226     | 6.2246     | 6.2265     | 6.2285     | 6.2305     | 6.2324     |
| <b>510</b>  | 2344       | 2364       | 2383       | 2403       | 2422       | 2442       | 2461       | 2480       | 2500       | 2519       |
| <b>520</b>  | 2538       | 2558       | 2577       | 2596       | 2615       | 2634       | 2653       | 2672       | 2691       | 2710       |
| <b>530</b>  | 2729       | 2748       | 2766       | 2785       | 2804       | 2823       | 2841       | 2860       | 2879       | 2897       |
| <b>540</b>  | 2916       | 2934       | 2953       | 2971       | 2989       | 3008       | 3026       | 3044       | 3063       | 3081       |
| <b>550</b>  | 3099       | 3117       | 3135       | 3154       | 3172       | 3190       | 3208       | 3226       | 3244       | 3261       |
| <b>560</b>  | 3279       | 3297       | 3315       | 3333       | 3351       | 3368       | 3386       | 3404       | 3421       | 3439       |
| <b>570</b>  | 3456       | 3474       | 3491       | 3509       | 3526       | 3544       | 3561       | 3578       | 3596       | 3613       |
| <b>580</b>  | 3630       | 3648       | 3665       | 3682       | 3699       | 3716       | 3733       | 3750       | 3767       | 3784       |
| <b>590</b>  | 3801       | 3818       | 3835       | 3852       | 3869       | 3886       | 3902       | 3919       | 3936       | 3953       |
| <b>600</b>  | 3969       | 3986       | 4003       | 4019       | 4036       | 4052       | 4069       | 4085       | 4102       | 4118       |
| <b>610</b>  | 4135       | 4151       | 4167       | 4184       | 4200       | 4216       | 4232       | 4249       | 4265       | 4281       |
| <b>620</b>  | 4297       | 4313       | 4329       | 4345       | 4362       | 4378       | 4394       | 4409       | 4425       | 4441       |
| <b>630</b>  | 4457       | 4473       | 4489       | 4505       | 4520       | 4536       | 4552       | 4568       | 4583       | 4599       |
| <b>640</b>  | 4615       | 4630       | 4646       | 4661       | 4677       | 4693       | 4708       | 4723       | 4739       | 4754       |
| <b>650</b>  | 4770       | 4785       | 4800       | 4816       | 4831       | 4846       | 4862       | 4877       | 4892       | 4907       |
| <b>660</b>  | 4922       | 4938       | 4953       | 4968       | 4983       | 4998       | 5013       | 5028       | 5043       | 5058       |
| <b>670</b>  | 5073       | 5088       | 5103       | 5117       | 5132       | 5147       | 5162       | 5177       | 5191       | 5206       |
| <b>680</b>  | 5221       | 5236       | 5250       | 5265       | 5280       | 5294       | 5309       | 5323       | 5338       | 5352       |
| <b>690</b>  | 5367       | 5381       | 5396       | 5410       | 5425       | 5439       | 5453       | 5468       | 5482       | 5497       |
| <b>700</b>  | 5511       | 5525       | 5539       | 5554       | 5568       | 5582       | 5596       | 5610       | 5624       | 5639       |
| <b>710</b>  | 5653       | 5667       | 5681       | 5695       | 5709       | 5723       | 5737       | 5751       | 5765       | 5779       |
| <b>720</b>  | 5793       | 5806       | 5820       | 5834       | 5848       | 5862       | 5876       | 5889       | 5903       | 5917       |
| <b>730</b>  | 5930       | 5944       | 5958       | 5971       | 5985       | 5999       | 6012       | 6026       | 6039       | 6053       |
| <b>740</b>  | 6067       | 6080       | 6093       | 6107       | 6120       | 6134       | 6147       | 6161       | 6174       | 6187       |
| <b>750</b>  | 6201       | 6214       | 6227       | 6241       | 6254       | 6267       | 6280       | 6294       | 6307       | 6320       |
| <b>760</b>  | 6333       | 6346       | 6359       | 6373       | 6386       | 6399       | 6412       | 6425       | 6438       | 6451       |
| <b>770</b>  | 6464       | 6477       | 6490       | 6503       | 6516       | 6529       | 6542       | 6554       | 6567       | 6580       |
| <b>780</b>  | 6593       | 6606       | 6619       | 6631       | 6644       | 6657       | 6670       | 6682       | 6695       | 6708       |
| <b>790</b>  | 6720       | 6733       | 6746       | 6758       | 6771       | 6783       | 6796       | 6809       | 6821       | 6834       |
| <b>800</b>  | 6846       | 6859       | 6871       | 6884       | 6896       | 6908       | 6921       | 6933       | 6946       | 6958       |
| <b>810</b>  | 6970       | 6983       | 6995       | 7007       | 7020       | 7032       | 7044       | 7056       | 7069       | 7081       |
| <b>820</b>  | 7093       | 7105       | 7117       | 7130       | 7142       | 7154       | 7166       | 7178       | 7190       | 7202       |
| <b>830</b>  | 7214       | 7226       | 7238       | 7250       | 7262       | 7274       | 7286       | 7298       | 7310       | 7322       |
| <b>840</b>  | 7334       | 7346       | 7358       | 7370       | 7382       | 7393       | 7405       | 7417       | 7429       | 7441       |
| <b>850</b>  | 7452       | 7464       | 7476       | 7488       | 7499       | 7511       | 7523       | 7534       | 7546       | 7558       |
| <b>860</b>  | 7569       | 7581       | 7593       | 7604       | 7616       | 7627       | 7639       | 7650       | 7662       | 7673       |
| <b>870</b>  | 7685       | 7696       | 7708       | 7719       | 7731       | 7742       | 7754       | 7765       | 7776       | 7788       |
| <b>880</b>  | 7799       | 7811       | 7822       | 7833       | 7845       | 7856       | 7867       | 7878       | 7890       | 7901       |
| <b>890</b>  | 7912       | 7923       | 7935       | 7946       | 7957       | 7968       | 7979       | 7991       | 8002       | 8013       |
| <b>900</b>  | 8024       | 8035       | 8046       | 8057       | 8068       | 8079       | 8090       | 8101       | 8112       | 8123       |
| <b>910</b>  | 8134       | 8145       | 8156       | 8167       | 8178       | 8189       | 8200       | 8211       | 8222       | 8233       |
| <b>920</b>  | 8244       | 8255       | 8265       | 8276       | 8287       | 8298       | 8309       | 8320       | 8330       | 8341       |
| <b>930</b>  | 8352       | 8363       | 8373       | 8384       | 8395       | 8405       | 8416       | 8427       | 8437       | 8448       |
| <b>940</b>  | 8459       | 8469       | 8480       | 8491       | 8501       | 8512       | 8522       | 8533       | 8544       | 8554       |
| <b>950</b>  | 8565       | 8575       | 8586       | 8596       | 8607       | 8617       | 8628       | 8638       | 8648       | 8659       |
| <b>960</b>  | 8669       | 8680       | 8690       | 8701       | 8711       | 8721       | 8732       | 8742       | 8752       | 8763       |
| <b>970</b>  | 8773       | 8783       | 8794       | 8804       | 8814       | 8824       | 8835       | 8845       | 8855       | 8865       |
| <b>980</b>  | 8876       | 8886       | 8896       | 8906       | 8916       | 8926       | 8937       | 8947       | 8957       | 8967       |
| <b>990</b>  | 8977       | 8987       | 8997       | 9007       | 9017       | 9027       | 9037       | 9048       | 9058       | 9068       |
| <b>1000</b> | 9078       | 9088       | 9098       | 9108       | 9117       | 9127       | 9137       | 9147       | 9157       | 9167       |
| <b>N</b>    | <b>0.0</b> | <b>1.0</b> | <b>2.0</b> | <b>3.0</b> | <b>4.0</b> | <b>5.0</b> | <b>6.0</b> | <b>7.0</b> | <b>8.0</b> | <b>9.0</b> |

# Values of $e^x$ and $e^{-x}$

0.00—1.99

| x   | Function | 0.00   | 0.01   | 0.02   | 0.03   | 0.04   | 0.05   | 0.06   | 0.07   | 0.08   | 0.09   |
|-----|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 0.0 | $e^x$    | 1.0000 | 1.0101 | 1.0202 | 1.0305 | 1.0408 | 1.0513 | 1.0618 | 1.0725 | 1.0833 | 1.0942 |
|     | $e^{-x}$ | 1.0000 | 0.9900 | 0.9802 | 0.9704 | 0.9608 | 0.9512 | 0.9418 | 0.9324 | 0.9231 | 0.9139 |
| 0.1 | $e^x$    | 1.1052 | 1.1163 | 1.1275 | 1.1388 | 1.1503 | 1.1618 | 1.1735 | 1.1853 | 1.1972 | 1.2093 |
|     | $e^{-x}$ | 0.9048 | 0.8958 | 0.8869 | 0.8781 | 0.8694 | 0.8607 | 0.8521 | 0.8437 | 0.8353 | 0.8270 |
| 0.2 | $e^x$    | 1.2214 | 1.2337 | 1.2461 | 1.2586 | 1.2712 | 1.2840 | 1.2969 | 1.3100 | 1.3231 | 1.3364 |
|     | $e^{-x}$ | 0.8187 | 0.8106 | 0.8025 | 0.7945 | 0.7866 | 0.7788 | 0.7711 | 0.7634 | 0.7558 | 0.7483 |
| 0.3 | $e^x$    | 1.3499 | 1.3634 | 1.3771 | 1.3910 | 1.4049 | 1.4191 | 1.4333 | 1.4477 | 1.4623 | 1.4770 |
|     | $e^{-x}$ | 0.7408 | 0.7334 | 0.7261 | 0.7189 | 0.7118 | 0.7047 | 0.6977 | 0.6907 | 0.6839 | 0.6771 |
| 0.4 | $e^x$    | 1.4918 | 1.5068 | 1.5220 | 1.5373 | 1.5527 | 1.5683 | 1.5841 | 1.6000 | 1.6161 | 1.6323 |
|     | $e^{-x}$ | 0.6703 | 0.6637 | 0.6570 | 0.6505 | 0.6440 | 0.6376 | 0.6313 | 0.6250 | 0.6188 | 0.6126 |
| 0.5 | $e^x$    | 1.6487 | 1.6653 | 1.6820 | 1.6989 | 1.7160 | 1.7333 | 1.7507 | 1.7683 | 1.7860 | 1.8040 |
|     | $e^{-x}$ | 0.6065 | 0.6005 | 0.5945 | 0.5886 | 0.5827 | 0.5769 | 0.5712 | 0.5655 | 0.5599 | 0.5543 |
| 0.6 | $e^x$    | 1.8221 | 1.8404 | 1.8589 | 1.8776 | 1.8965 | 1.9155 | 1.9348 | 1.9542 | 1.9739 | 1.9939 |
|     | $e^{-x}$ | 0.5488 | 0.5434 | 0.5379 | 0.5326 | 0.5273 | 0.5220 | 0.5169 | 0.5117 | 0.5066 | 0.5017 |
| 0.7 | $e^x$    | 2.0138 | 2.0340 | 2.0544 | 2.0751 | 2.0959 | 2.1170 | 2.1383 | 2.1598 | 2.1815 | 2.2034 |
|     | $e^{-x}$ | 0.4966 | 0.4916 | 0.4868 | 0.4819 | 0.4771 | 0.4724 | 0.4677 | 0.4630 | 0.4584 | 0.4538 |
| 0.8 | $e^x$    | 2.2255 | 2.2479 | 2.2705 | 2.2933 | 2.3164 | 2.3396 | 2.3632 | 2.3869 | 2.4109 | 2.4351 |
|     | $e^{-x}$ | 0.4493 | 0.4449 | 0.4404 | 0.4360 | 0.4317 | 0.4274 | 0.4232 | 0.4190 | 0.4148 | 0.4107 |
| 0.9 | $e^x$    | 2.4596 | 2.4843 | 2.5093 | 2.5345 | 2.5600 | 2.5857 | 2.6117 | 2.6379 | 2.6645 | 2.6912 |
|     | $e^{-x}$ | 0.4066 | 0.4025 | 0.3985 | 0.3946 | 0.3906 | 0.3867 | 0.3829 | 0.3791 | 0.3753 | 0.3716 |
| 1.0 | $e^x$    | 2.7183 | 2.7456 | 2.7732 | 2.8011 | 2.8292 | 2.8577 | 2.8864 | 2.9154 | 2.9447 | 2.9743 |
|     | $e^{-x}$ | 0.3679 | 0.3642 | 0.3606 | 0.3570 | 0.3535 | 0.3499 | 0.3465 | 0.3430 | 0.3396 | 0.3362 |
| 1.1 | $e^x$    | 3.0042 | 3.0344 | 3.0649 | 3.0957 | 3.1268 | 3.1582 | 3.1899 | 3.2220 | 3.2544 | 3.2871 |
|     | $e^{-x}$ | 0.3329 | 0.3296 | 0.3263 | 0.3230 | 0.3198 | 0.3166 | 0.3135 | 0.3104 | 0.3073 | 0.3042 |
| 1.2 | $e^x$    | 3.3201 | 3.3535 | 3.3872 | 3.4212 | 3.4556 | 3.4903 | 3.5254 | 3.5609 | 3.5966 | 3.6328 |
|     | $e^{-x}$ | 0.3012 | 0.2982 | 0.2952 | 0.2923 | 0.2894 | 0.2865 | 0.2837 | 0.2808 | 0.2780 | 0.2753 |
| 1.3 | $e^x$    | 3.6693 | 3.7062 | 3.7434 | 3.7810 | 3.8190 | 3.8574 | 3.8962 | 3.9354 | 3.9749 | 4.0149 |
|     | $e^{-x}$ | 0.2725 | 0.2698 | 0.2671 | 0.2645 | 0.2618 | 0.2592 | 0.2567 | 0.2541 | 0.2516 | 0.2491 |
| 1.4 | $e^x$    | 4.0552 | 4.0960 | 4.1371 | 4.1787 | 4.2207 | 4.2631 | 4.3060 | 4.3492 | 4.3929 | 4.4371 |
|     | $e^{-x}$ | 0.2466 | 0.2441 | 0.2417 | 0.2393 | 0.2369 | 0.2346 | 0.2322 | 0.2299 | 0.2276 | 0.2254 |
| 1.5 | $e^x$    | 4.4817 | 4.5267 | 4.5722 | 4.6182 | 4.6646 | 4.7115 | 4.7588 | 4.8066 | 4.8550 | 4.9037 |
|     | $e^{-x}$ | 0.2231 | 0.2209 | 0.2187 | 0.2165 | 0.2144 | 0.2122 | 0.2101 | 0.2080 | 0.2060 | 0.2039 |
| 1.6 | $e^x$    | 4.9530 | 5.0028 | 5.0531 | 5.1039 | 5.1552 | 5.2070 | 5.2593 | 5.3122 | 5.3656 | 5.4195 |
|     | $e^{-x}$ | 0.2019 | 0.1999 | 0.1979 | 0.1959 | 0.1940 | 0.1920 | 0.1901 | 0.1882 | 0.1864 | 0.1845 |
| 1.7 | $e^x$    | 5.4739 | 5.5290 | 5.5845 | 5.6407 | 5.6973 | 5.7546 | 5.8124 | 5.8709 | 5.9299 | 5.9895 |
|     | $e^{-x}$ | 0.1827 | 0.1809 | 0.1791 | 0.1773 | 0.1755 | 0.1738 | 0.1720 | 0.1703 | 0.1686 | 0.1670 |
| 1.8 | $e^x$    | 6.0496 | 6.1104 | 6.1719 | 6.2339 | 6.2965 | 6.3598 | 6.4237 | 6.4883 | 6.5535 | 6.6194 |
|     | $e^{-x}$ | 0.1653 | 0.1637 | 0.1620 | 0.1604 | 0.1588 | 0.1572 | 0.1557 | 0.1541 | 0.1526 | 0.1511 |
| 1.9 | $e^x$    | 6.6859 | 6.7531 | 6.8210 | 6.8895 | 6.9588 | 7.0287 | 7.0993 | 7.1707 | 7.2427 | 7.3155 |
|     | $e^{-x}$ | 0.1496 | 0.1481 | 0.1466 | 0.1451 | 0.1437 | 0.1423 | 0.1409 | 0.1395 | 0.1381 | 0.1367 |

Values of  $e^x$  and  $e^{-x}$

2.00-3.99

| x   | Function | 0.00   | 0.01   | 0.02   | 0.03   | 0.04   | 0.05   | 0.06   | 0.07   | 0.08   | 0.09   |
|-----|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 2.0 | $e^x$    | 7.3891 | 7.4633 | 7.5383 | 7.6141 | 7.6906 | 7.7679 | 7.8460 | 7.9248 | 8.0045 | 8.0849 |
|     | $e^{-x}$ | 0.1353 | 0.1340 | 0.1327 | 0.1313 | 0.1300 | 0.1287 | 0.1275 | 0.1262 | 0.1249 | 0.1237 |
| 2.1 | $e^x$    | 8.1662 | 8.2482 | 8.3311 | 8.4149 | 8.4994 | 8.5849 | 8.6711 | 8.7583 | 8.8463 | 8.9352 |
|     | $e^{-x}$ | 0.1225 | 0.1212 | 0.1200 | 0.1188 | 0.1177 | 0.1165 | 0.1153 | 0.1142 | 0.1130 | 0.1119 |
| 2.2 | $e^x$    | 9.0250 | 9.1157 | 9.2073 | 9.2999 | 9.3933 | 9.4877 | 9.5831 | 9.6794 | 9.7767 | 9.8749 |
|     | $e^{-x}$ | 0.1108 | 0.1097 | 0.1086 | 0.1075 | 0.1065 | 0.1054 | 0.1044 | 0.1033 | 0.1023 | 0.1013 |
| 2.3 | $e^x$    | 9.9742 | 10.074 | 10.176 | 10.278 | 10.381 | 10.486 | 10.591 | 10.697 | 10.805 | 10.913 |
|     | $e^{-x}$ | 0.1003 | 0.0993 | 0.0983 | 0.0973 | 0.0963 | 0.0954 | 0.0944 | 0.0935 | 0.0926 | 0.0916 |
| 2.4 | $e^x$    | 11.023 | 11.134 | 11.246 | 11.359 | 11.473 | 11.588 | 11.705 | 11.822 | 11.941 | 12.061 |
|     | $e^{-x}$ | 0.0907 | 0.0898 | 0.0889 | 0.0880 | 0.0872 | 0.0863 | 0.0854 | 0.0846 | 0.0837 | 0.0829 |
| 2.5 | $e^x$    | 12.182 | 12.305 | 12.429 | 12.554 | 12.680 | 12.807 | 12.936 | 13.066 | 13.197 | 13.330 |
|     | $e^{-x}$ | 0.0821 | 0.0813 | 0.0805 | 0.0797 | 0.0789 | 0.0781 | 0.0773 | 0.0765 | 0.0758 | 0.0750 |
| 2.6 | $e^x$    | 13.464 | 13.599 | 13.736 | 13.874 | 14.013 | 14.154 | 14.296 | 14.440 | 14.585 | 14.732 |
|     | $e^{-x}$ | 0.0743 | 0.0735 | 0.0728 | 0.0721 | 0.0714 | 0.0707 | 0.0699 | 0.0693 | 0.0686 | 0.0679 |
| 2.7 | $e^x$    | 14.880 | 15.029 | 15.180 | 15.333 | 15.487 | 15.643 | 15.800 | 15.959 | 16.119 | 16.281 |
|     | $e^{-x}$ | 0.0672 | 0.0665 | 0.0659 | 0.0652 | 0.0646 | 0.0639 | 0.0633 | 0.0627 | 0.0620 | 0.0614 |
| 2.8 | $e^x$    | 16.445 | 16.610 | 16.777 | 16.945 | 17.116 | 17.288 | 17.462 | 17.637 | 17.814 | 17.993 |
|     | $e^{-x}$ | 0.0608 | 0.0602 | 0.0596 | 0.0590 | 0.0584 | 0.0578 | 0.0573 | 0.0567 | 0.0561 | 0.0556 |
| 2.9 | $e^x$    | 18.174 | 18.357 | 18.541 | 18.728 | 18.916 | 19.106 | 19.298 | 19.492 | 19.688 | 19.886 |
|     | $e^{-x}$ | 0.0550 | 0.0545 | 0.0539 | 0.0534 | 0.0529 | 0.0523 | 0.0518 | 0.0513 | 0.0508 | 0.0503 |
| 3.0 | $e^x$    | 20.086 | 20.287 | 20.491 | 20.697 | 20.905 | 21.115 | 21.328 | 21.542 | 21.758 | 21.977 |
|     | $e^{-x}$ | 0.0498 | 0.0493 | 0.0488 | 0.0483 | 0.0478 | 0.0474 | 0.0469 | 0.0464 | 0.0460 | 0.0455 |
| 3.1 | $e^x$    | 22.198 | 22.421 | 22.646 | 22.874 | 23.104 | 23.336 | 23.571 | 23.807 | 24.047 | 24.288 |
|     | $e^{-x}$ | 0.0450 | 0.0446 | 0.0442 | 0.0437 | 0.0433 | 0.0429 | 0.0424 | 0.0420 | 0.0416 | 0.0412 |
| 3.2 | $e^x$    | 24.533 | 24.779 | 25.028 | 25.280 | 25.534 | 25.790 | 26.050 | 26.311 | 26.576 | 26.843 |
|     | $e^{-x}$ | 0.0408 | 0.0404 | 0.0400 | 0.0396 | 0.0392 | 0.0388 | 0.0384 | 0.0380 | 0.0376 | 0.0373 |
| 3.3 | $e^x$    | 27.113 | 27.385 | 27.660 | 27.938 | 28.219 | 28.503 | 28.789 | 29.079 | 29.371 | 29.666 |
|     | $e^{-x}$ | 0.0369 | 0.0365 | 0.0362 | 0.0358 | 0.0354 | 0.0351 | 0.0347 | 0.0344 | 0.0340 | 0.0337 |
| 3.4 | $e^x$    | 29.964 | 30.265 | 30.569 | 30.877 | 31.187 | 31.500 | 31.817 | 32.137 | 32.460 | 32.786 |
|     | $e^{-x}$ | 0.0334 | 0.0330 | 0.0327 | 0.0324 | 0.0321 | 0.0317 | 0.0314 | 0.0311 | 0.0308 | 0.0305 |
| 3.5 | $e^x$    | 33.115 | 33.448 | 33.784 | 34.124 | 34.467 | 34.813 | 35.163 | 35.517 | 35.874 | 36.234 |
|     | $e^{-x}$ | 0.0302 | 0.0299 | 0.0296 | 0.0293 | 0.0290 | 0.0287 | 0.0284 | 0.0282 | 0.0279 | 0.0276 |
| 3.6 | $e^x$    | 36.598 | 36.966 | 37.338 | 37.713 | 38.092 | 38.475 | 38.861 | 39.252 | 39.646 | 40.045 |
|     | $e^{-x}$ | 0.0273 | 0.0271 | 0.0268 | 0.0265 | 0.0263 | 0.0260 | 0.0257 | 0.0255 | 0.0252 | 0.0250 |
| 3.7 | $e^x$    | 40.447 | 40.854 | 41.264 | 41.679 | 42.098 | 42.521 | 42.948 | 43.380 | 43.816 | 44.256 |
|     | $e^{-x}$ | 0.0247 | 0.0245 | 0.0242 | 0.0240 | 0.0238 | 0.0235 | 0.0233 | 0.0231 | 0.0228 | 0.0226 |
| 3.8 | $e^x$    | 44.701 | 45.150 | 45.604 | 46.063 | 46.525 | 46.993 | 47.465 | 47.942 | 48.424 | 48.911 |
|     | $e^{-x}$ | 0.0224 | 0.0221 | 0.0219 | 0.0217 | 0.0215 | 0.0213 | 0.0211 | 0.0209 | 0.0207 | 0.0204 |
| 3.9 | $e^x$    | 49.402 | 49.899 | 50.400 | 50.907 | 51.419 | 51.935 | 52.457 | 52.985 | 53.517 | 54.055 |
|     | $e^{-x}$ | 0.0202 | 0.0200 | 0.0198 | 0.0196 | 0.0195 | 0.0193 | 0.0191 | 0.0189 | 0.0187 | 0.0185 |

Values of  $e^x$  and  $e^{-x}$

4.00-5.99

| x   | Function | 0.00   | 0.01   | 0.02   | 0.03   | 0.04   | 0.05   | 0.06   | 0.07   | 0.08   | 0.09   |
|-----|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 4.0 | $e^x$    | 54.598 | 55.147 | 55.701 | 56.261 | 56.826 | 57.397 | 57.974 | 58.557 | 59.145 | 59.740 |
|     | $e^{-x}$ | 0.0183 | 0.0181 | 0.0180 | 0.0178 | 0.0176 | 0.0174 | 0.0172 | 0.0171 | 0.0169 | 0.0167 |
| 4.1 | $e^x$    | 60.340 | 60.947 | 61.559 | 62.178 | 62.803 | 63.434 | 64.072 | 64.715 | 65.366 | 66.023 |
|     | $e^{-x}$ | 0.0166 | 0.0164 | 0.0162 | 0.0161 | 0.0159 | 0.0158 | 0.0156 | 0.0155 | 0.0153 | 0.0151 |
| 4.2 | $e^x$    | 66.686 | 67.357 | 68.033 | 68.717 | 69.408 | 70.105 | 70.810 | 71.522 | 72.240 | 72.966 |
|     | $e^{-x}$ | 0.0150 | 0.0148 | 0.0147 | 0.0146 | 0.0144 | 0.0143 | 0.0141 | 0.0140 | 0.0138 | 0.0137 |
| 4.3 | $e^x$    | 73.700 | 74.440 | 75.189 | 75.944 | 76.708 | 77.478 | 78.257 | 79.044 | 79.838 | 80.640 |
|     | $e^{-x}$ | 0.0136 | 0.0134 | 0.0133 | 0.0132 | 0.0130 | 0.0129 | 0.0128 | 0.0127 | 0.0125 | 0.0124 |
| 4.4 | $e^x$    | 81.451 | 82.269 | 83.096 | 83.931 | 84.775 | 85.627 | 86.488 | 87.357 | 88.235 | 89.121 |
|     | $e^{-x}$ | 0.0123 | 0.0122 | 0.0120 | 0.0119 | 0.0118 | 0.0117 | 0.0116 | 0.0114 | 0.0113 | 0.0112 |
| 4.5 | $e^x$    | 90.017 | 90.922 | 91.836 | 92.759 | 93.691 | 94.632 | 95.583 | 96.544 | 97.514 | 98.494 |
|     | $e^{-x}$ | 0.0111 | 0.0110 | 0.0109 | 0.0108 | 0.0107 | 0.0106 | 0.0105 | 0.0104 | 0.0103 | 0.0102 |
| 4.6 | $e^x$    | 99.484 | 100.48 | 101.49 | 102.51 | 103.54 | 104.58 | 105.64 | 106.70 | 107.77 | 108.85 |
|     | $e^{-x}$ | 0.0101 | 0.0100 | 0.0099 | 0.0098 | 0.0097 | 0.0096 | 0.0095 | 0.0094 | 0.0093 | 0.0092 |
| 4.7 | $e^x$    | 109.95 | 111.05 | 112.17 | 113.30 | 114.43 | 115.58 | 116.75 | 117.92 | 119.10 | 120.30 |
|     | $e^{-x}$ | 0.0091 | 0.0090 | 0.0089 | 0.0088 | 0.0087 | 0.0087 | 0.0086 | 0.0085 | 0.0084 | 0.0083 |
| 4.8 | $e^x$    | 121.51 | 122.73 | 123.97 | 125.21 | 126.47 | 127.74 | 129.02 | 130.32 | 131.63 | 132.95 |
|     | $e^{-x}$ | 0.0082 | 0.0081 | 0.0081 | 0.0080 | 0.0079 | 0.0078 | 0.0078 | 0.0077 | 0.0076 | 0.0075 |
| 4.9 | $e^x$    | 134.29 | 135.64 | 137.00 | 138.38 | 139.77 | 141.17 | 142.59 | 144.03 | 145.47 | 146.94 |
|     | $e^{-x}$ | 0.0074 | 0.0074 | 0.0073 | 0.0072 | 0.0072 | 0.0071 | 0.0070 | 0.0069 | 0.0069 | 0.0068 |
| 5.0 | $e^x$    | 148.41 | 149.90 | 151.41 | 152.93 | 154.47 | 156.02 | 157.59 | 159.17 | 160.77 | 162.39 |
|     | $e^{-x}$ | 0.0067 | 0.0067 | 0.0066 | 0.0065 | 0.0065 | 0.0064 | 0.0063 | 0.0063 | 0.0062 | 0.0062 |
| 5.1 | $e^x$    | 164.02 | 165.67 | 167.34 | 169.02 | 170.72 | 172.43 | 174.16 | 175.91 | 177.68 | 179.47 |
|     | $e^{-x}$ | 0.0061 | 0.0060 | 0.0060 | 0.0059 | 0.0059 | 0.0058 | 0.0057 | 0.0057 | 0.0056 | 0.0056 |
| 5.2 | $e^x$    | 181.27 | 183.09 | 184.93 | 186.79 | 188.67 | 190.57 | 192.48 | 194.42 | 196.37 | 198.34 |
|     | $e^{-x}$ | 0.0055 | 0.0055 | 0.0054 | 0.0054 | 0.0053 | 0.0052 | 0.0052 | 0.0051 | 0.0051 | 0.0050 |
| 5.3 | $e^x$    | 200.34 | 202.35 | 204.38 | 206.44 | 208.51 | 210.61 | 212.72 | 214.86 | 217.02 | 219.20 |
|     | $e^{-x}$ | 0.0050 | 0.0049 | 0.0049 | 0.0048 | 0.0048 | 0.0047 | 0.0047 | 0.0047 | 0.0046 | 0.0046 |
| 5.4 | $e^x$    | 221.41 | 223.63 | 225.88 | 228.15 | 230.44 | 232.76 | 235.10 | 237.46 | 239.85 | 242.26 |
|     | $e^{-x}$ | 0.0045 | 0.0045 | 0.0044 | 0.0044 | 0.0043 | 0.0043 | 0.0043 | 0.0042 | 0.0042 | 0.0041 |
| 5.5 | $e^x$    | 244.69 | 247.15 | 249.64 | 252.14 | 254.68 | 257.24 | 259.82 | 262.43 | 265.07 | 267.74 |
|     | $e^{-x}$ | 0.0041 | 0.0040 | 0.0040 | 0.0040 | 0.0039 | 0.0039 | 0.0038 | 0.0038 | 0.0038 | 0.0037 |
| 5.6 | $e^x$    | 270.43 | 273.14 | 275.89 | 278.66 | 281.46 | 284.29 | 287.15 | 290.03 | 292.95 | 295.89 |
|     | $e^{-x}$ | 0.0037 | 0.0037 | 0.0036 | 0.0036 | 0.0036 | 0.0035 | 0.0035 | 0.0034 | 0.0034 | 0.0034 |
| 5.7 | $e^x$    | 298.87 | 301.87 | 304.90 | 307.97 | 311.06 | 314.19 | 317.35 | 320.54 | 323.76 | 327.01 |
|     | $e^{-x}$ | 0.0033 | 0.0033 | 0.0033 | 0.0032 | 0.0032 | 0.0032 | 0.0032 | 0.0031 | 0.0031 | 0.0031 |
| 5.8 | $e^x$    | 330.30 | 333.62 | 336.97 | 340.36 | 343.78 | 347.23 | 350.72 | 354.25 | 357.81 | 361.41 |
|     | $e^{-x}$ | 0.0030 | 0.0030 | 0.0030 | 0.0029 | 0.0029 | 0.0029 | 0.0029 | 0.0028 | 0.0028 | 0.0028 |
| 5.9 | $e^x$    | 365.04 | 368.71 | 372.41 | 376.15 | 379.93 | 383.75 | 387.61 | 391.51 | 395.44 | 399.41 |
|     | $e^{-x}$ | 0.0027 | 0.0027 | 0.0027 | 0.0027 | 0.0026 | 0.0026 | 0.0026 | 0.0026 | 0.0025 | 0.0025 |



# Natural Sines, Cosines and Tangents

0°-14.9°

| Degs. | Function | 0.0°   | 0.1°   | 0.2°   | 0.3°   | 0.4°   | 0.5°   | 0.6°   | 0.7°   | 0.8°   | 0.9°   |
|-------|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 0     | sin      | 0.0000 | 0.0017 | 0.0035 | 0.0052 | 0.0070 | 0.0087 | 0.0105 | 0.0122 | 0.0140 | 0.0157 |
|       | cos      | 1.0000 | 1.0000 | 1.0000 | 1.0000 | 1.0000 | 1.0000 | 0.9999 | 0.9999 | 0.9999 | 0.9999 |
|       | tan      | 0.0000 | 0.0017 | 0.0035 | 0.0052 | 0.0070 | 0.0087 | 0.0105 | 0.0122 | 0.0140 | 0.0157 |
| 1     | sin      | 0.0175 | 0.0192 | 0.0209 | 0.0227 | 0.0244 | 0.0262 | 0.0279 | 0.0297 | 0.0314 | 0.0332 |
|       | cos      | 0.9998 | 0.9998 | 0.9998 | 0.9997 | 0.9997 | 0.9997 | 0.9996 | 0.9996 | 0.9995 | 0.9995 |
|       | tan      | 0.0175 | 0.0192 | 0.0209 | 0.0227 | 0.0244 | 0.0262 | 0.0279 | 0.0297 | 0.0314 | 0.0332 |
| 2     | sin      | 0.0349 | 0.0366 | 0.0384 | 0.0401 | 0.0419 | 0.0436 | 0.0454 | 0.0471 | 0.0488 | 0.0506 |
|       | cos      | 0.9994 | 0.9993 | 0.9993 | 0.9992 | 0.9991 | 0.9990 | 0.9990 | 0.9989 | 0.9988 | 0.9987 |
|       | tan      | 0.0349 | 0.0367 | 0.0384 | 0.0402 | 0.0419 | 0.0437 | 0.0454 | 0.0472 | 0.0489 | 0.0507 |
| 3     | sin      | 0.0523 | 0.0541 | 0.0558 | 0.0576 | 0.0593 | 0.0610 | 0.0628 | 0.0645 | 0.0663 | 0.0680 |
|       | cos      | 0.9986 | 0.9985 | 0.9984 | 0.9983 | 0.9982 | 0.9981 | 0.9980 | 0.9979 | 0.9978 | 0.9977 |
|       | tan      | 0.0524 | 0.0542 | 0.0559 | 0.0577 | 0.0594 | 0.0612 | 0.0629 | 0.0647 | 0.0664 | 0.0682 |
| 4     | sin      | 0.0698 | 0.0715 | 0.0732 | 0.0750 | 0.0767 | 0.0785 | 0.0802 | 0.0819 | 0.0837 | 0.0854 |
|       | cos      | 0.9976 | 0.9974 | 0.9973 | 0.9972 | 0.9971 | 0.9969 | 0.9968 | 0.9966 | 0.9965 | 0.9963 |
|       | tan      | 0.0699 | 0.0717 | 0.0734 | 0.0752 | 0.0769 | 0.0787 | 0.0805 | 0.0822 | 0.0840 | 0.0857 |
| 5     | sin      | 0.0872 | 0.0889 | 0.0906 | 0.0924 | 0.0941 | 0.0958 | 0.0976 | 0.0993 | 0.1011 | 0.1028 |
|       | cos      | 0.9962 | 0.9960 | 0.9959 | 0.9957 | 0.9956 | 0.9954 | 0.9952 | 0.9951 | 0.9949 | 0.9947 |
|       | tan      | 0.0875 | 0.0892 | 0.0910 | 0.0928 | 0.0945 | 0.0963 | 0.0981 | 0.0998 | 0.1016 | 0.1033 |
| 6     | sin      | 0.1045 | 0.1063 | 0.1080 | 0.1097 | 0.1115 | 0.1132 | 0.1149 | 0.1167 | 0.1184 | 0.1201 |
|       | cos      | 0.9945 | 0.9943 | 0.9942 | 0.9940 | 0.9938 | 0.9936 | 0.9934 | 0.9932 | 0.9930 | 0.9928 |
|       | tan      | 0.1051 | 0.1069 | 0.1086 | 0.1104 | 0.1122 | 0.1139 | 0.1157 | 0.1175 | 0.1192 | 0.1210 |
| 7     | sin      | 0.1219 | 0.1236 | 0.1253 | 0.1271 | 0.1288 | 0.1305 | 0.1323 | 0.1340 | 0.1357 | 0.1374 |
|       | cos      | 0.9925 | 0.9923 | 0.9921 | 0.9919 | 0.9917 | 0.9914 | 0.9912 | 0.9910 | 0.9907 | 0.9905 |
|       | tan      | 0.1228 | 0.1246 | 0.1263 | 0.1281 | 0.1299 | 0.1317 | 0.1334 | 0.1352 | 0.1370 | 0.1388 |
| 8     | sin      | 0.1392 | 0.1409 | 0.1426 | 0.1444 | 0.1461 | 0.1478 | 0.1495 | 0.1513 | 0.1530 | 0.1547 |
|       | cos      | 0.9903 | 0.9900 | 0.9898 | 0.9895 | 0.9893 | 0.9890 | 0.9888 | 0.9885 | 0.9882 | 0.9880 |
|       | tan      | 0.1405 | 0.1423 | 0.1441 | 0.1459 | 0.1477 | 0.1495 | 0.1512 | 0.1530 | 0.1548 | 0.1566 |
| 9     | sin      | 0.1564 | 0.1582 | 0.1599 | 0.1616 | 0.1633 | 0.1650 | 0.1668 | 0.1685 | 0.1702 | 0.1719 |
|       | cos      | 0.9877 | 0.9874 | 0.9871 | 0.9869 | 0.9866 | 0.9863 | 0.9860 | 0.9857 | 0.9854 | 0.9851 |
|       | tan      | 0.1584 | 0.1602 | 0.1620 | 0.1638 | 0.1655 | 0.1673 | 0.1691 | 0.1709 | 0.1727 | 0.1745 |
| 10    | sin      | 0.1736 | 0.1754 | 0.1771 | 0.1788 | 0.1805 | 0.1822 | 0.1840 | 0.1857 | 0.1874 | 0.1891 |
|       | cos      | 0.9848 | 0.9845 | 0.9842 | 0.9839 | 0.9836 | 0.9833 | 0.9829 | 0.9826 | 0.9823 | 0.9820 |
|       | tan      | 0.1763 | 0.1781 | 0.1799 | 0.1817 | 0.1835 | 0.1853 | 0.1871 | 0.1890 | 0.1908 | 0.1926 |
| 11    | sin      | 0.1908 | 0.1925 | 0.1942 | 0.1959 | 0.1977 | 0.1994 | 0.2011 | 0.2028 | 0.2045 | 0.2062 |
|       | cos      | 0.9816 | 0.9813 | 0.9810 | 0.9806 | 0.9803 | 0.9799 | 0.9796 | 0.9792 | 0.9789 | 0.9785 |
|       | tan      | 0.1944 | 0.1962 | 0.1980 | 0.1998 | 0.2016 | 0.2035 | 0.2053 | 0.2071 | 0.2089 | 0.2107 |
| 12    | sin      | 0.2079 | 0.2096 | 0.2113 | 0.2130 | 0.2147 | 0.2164 | 0.2181 | 0.2198 | 0.2215 | 0.2232 |
|       | cos      | 0.9781 | 0.9778 | 0.9774 | 0.9770 | 0.9767 | 0.9763 | 0.9759 | 0.9755 | 0.9751 | 0.9748 |
|       | tan      | 0.2126 | 0.2144 | 0.2162 | 0.2180 | 0.2199 | 0.2217 | 0.2235 | 0.2254 | 0.2272 | 0.2290 |
| 13    | sin      | 0.2250 | 0.2267 | 0.2284 | 0.2300 | 0.2318 | 0.2334 | 0.2351 | 0.2368 | 0.2385 | 0.2402 |
|       | cos      | 0.9744 | 0.9740 | 0.9736 | 0.9732 | 0.9728 | 0.9724 | 0.9720 | 0.9715 | 0.9711 | 0.9707 |
|       | tan      | 0.2309 | 0.2327 | 0.2345 | 0.2364 | 0.2382 | 0.2401 | 0.2419 | 0.2438 | 0.2456 | 0.2475 |
| 14    | sin      | 0.2419 | 0.2436 | 0.2453 | 0.2470 | 0.2487 | 0.2504 | 0.2521 | 0.2538 | 0.2554 | 0.2571 |
|       | cos      | 0.9703 | 0.9699 | 0.9694 | 0.9690 | 0.9686 | 0.9681 | 0.9677 | 0.9673 | 0.9668 | 0.9664 |
|       | tan      | 0.2493 | 0.2512 | 0.2530 | 0.2549 | 0.2568 | 0.2586 | 0.2605 | 0.2623 | 0.2642 | 0.2661 |
| Degs. | Function | 0'     | 6'     | 12'    | 18'    | 24'    | 30'    | 36'    | 42'    | 48'    | 54'    |

# Natural Sines, Cosines and Tangents

15°-29.9°

| Degs. | Function | 0.0°   | 0.1°   | 0.2°   | 0.3°   | 0.4°   | 0.5°   | 0.6°   | 0.7°   | 0.8°   | 0.9°   |
|-------|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 15    | sin      | 0.2588 | 0.2605 | 0.2622 | 0.2639 | 0.2656 | 0.2672 | 0.2689 | 0.2706 | 0.2723 | 0.2740 |
|       | cos      | 0.9659 | 0.9655 | 0.9650 | 0.9646 | 0.9641 | 0.9636 | 0.9632 | 0.9627 | 0.9622 | 0.9617 |
|       | tan      | 0.2679 | 0.2698 | 0.2717 | 0.2736 | 0.2754 | 0.2773 | 0.2792 | 0.2811 | 0.2830 | 0.2849 |
| 16    | sin      | 0.2756 | 0.2773 | 0.2790 | 0.2807 | 0.2823 | 0.2840 | 0.2857 | 0.2874 | 0.2890 | 0.2907 |
|       | cos      | 0.9613 | 0.9608 | 0.9603 | 0.9598 | 0.9593 | 0.9588 | 0.9583 | 0.9578 | 0.9573 | 0.9568 |
|       | tan      | 0.2867 | 0.2886 | 0.2905 | 0.2924 | 0.2943 | 0.2962 | 0.2981 | 0.3000 | 0.3019 | 0.3038 |
| 17    | sin      | 0.2924 | 0.2940 | 0.2957 | 0.2974 | 0.2990 | 0.3007 | 0.3024 | 0.3040 | 0.3057 | 0.3074 |
|       | cos      | 0.9563 | 0.9558 | 0.9553 | 0.9548 | 0.9542 | 0.9537 | 0.9532 | 0.9527 | 0.9521 | 0.9516 |
|       | tan      | 0.3057 | 0.3076 | 0.3096 | 0.3115 | 0.3134 | 0.3153 | 0.3172 | 0.3191 | 0.3211 | 0.3230 |
| 18    | sin      | 0.3090 | 0.3107 | 0.3123 | 0.3140 | 0.3156 | 0.3173 | 0.3190 | 0.3206 | 0.3223 | 0.3239 |
|       | cos      | 0.9511 | 0.9505 | 0.9500 | 0.9494 | 0.9489 | 0.9483 | 0.9478 | 0.9472 | 0.9466 | 0.9461 |
|       | tan      | 0.3249 | 0.3269 | 0.3288 | 0.3307 | 0.3327 | 0.3346 | 0.3365 | 0.3385 | 0.3404 | 0.3424 |
| 19    | sin      | 0.3256 | 0.3272 | 0.3289 | 0.3305 | 0.3322 | 0.3338 | 0.3355 | 0.3371 | 0.3387 | 0.3404 |
|       | cos      | 0.9455 | 0.9449 | 0.9444 | 0.9438 | 0.9432 | 0.9426 | 0.9421 | 0.9415 | 0.9409 | 0.9403 |
|       | tan      | 0.3443 | 0.3463 | 0.3482 | 0.3502 | 0.3522 | 0.3541 | 0.3561 | 0.3581 | 0.3600 | 0.3620 |
| 20    | sin      | 0.3420 | 0.3437 | 0.3453 | 0.3469 | 0.3486 | 0.3502 | 0.3518 | 0.3535 | 0.3551 | 0.3567 |
|       | cos      | 0.9397 | 0.9391 | 0.9385 | 0.9379 | 0.9373 | 0.9367 | 0.9361 | 0.9354 | 0.9348 | 0.9342 |
|       | tan      | 0.3640 | 0.3659 | 0.3679 | 0.3699 | 0.3719 | 0.3739 | 0.3759 | 0.3779 | 0.3799 | 0.3819 |
| 21    | sin      | 0.3584 | 0.3600 | 0.3616 | 0.3633 | 0.3649 | 0.3665 | 0.3681 | 0.3697 | 0.3714 | 0.3730 |
|       | cos      | 0.9336 | 0.9330 | 0.9323 | 0.9317 | 0.9311 | 0.9304 | 0.9298 | 0.9291 | 0.9285 | 0.9278 |
|       | tan      | 0.3839 | 0.3859 | 0.3879 | 0.3899 | 0.3919 | 0.3939 | 0.3959 | 0.3979 | 0.4000 | 0.4020 |
| 22    | sin      | 0.3746 | 0.3762 | 0.3778 | 0.3795 | 0.3811 | 0.3827 | 0.3843 | 0.3859 | 0.3875 | 0.3891 |
|       | cos      | 0.9272 | 0.9265 | 0.9259 | 0.9252 | 0.9245 | 0.9239 | 0.9232 | 0.9225 | 0.9219 | 0.9212 |
|       | tan      | 0.4040 | 0.4061 | 0.4081 | 0.4101 | 0.4122 | 0.4142 | 0.4163 | 0.4183 | 0.4204 | 0.4224 |
| 23    | sin      | 0.3907 | 0.3923 | 0.3939 | 0.3955 | 0.3971 | 0.3987 | 0.4003 | 0.4019 | 0.4035 | 0.4051 |
|       | cos      | 0.9205 | 0.9198 | 0.9191 | 0.9184 | 0.9178 | 0.9171 | 0.9164 | 0.9157 | 0.9150 | 0.9143 |
|       | tan      | 0.4245 | 0.4265 | 0.4286 | 0.4307 | 0.4327 | 0.4348 | 0.4369 | 0.4390 | 0.4411 | 0.4431 |
| 24    | sin      | 0.4067 | 0.4083 | 0.4099 | 0.4115 | 0.4131 | 0.4147 | 0.4163 | 0.4179 | 0.4195 | 0.4210 |
|       | cos      | 0.9135 | 0.9128 | 0.9121 | 0.9114 | 0.9107 | 0.9100 | 0.9092 | 0.9085 | 0.9078 | 0.9070 |
|       | tan      | 0.4452 | 0.4473 | 0.4494 | 0.4515 | 0.4536 | 0.4557 | 0.4578 | 0.4599 | 0.4621 | 0.4642 |
| 25    | sin      | 0.4226 | 0.4242 | 0.4258 | 0.4274 | 0.4289 | 0.4305 | 0.4321 | 0.4337 | 0.4352 | 0.4368 |
|       | cos      | 0.9063 | 0.9056 | 0.9048 | 0.9041 | 0.9033 | 0.9026 | 0.9018 | 0.9011 | 0.9003 | 0.8996 |
|       | tan      | 0.4663 | 0.4684 | 0.4706 | 0.4727 | 0.4748 | 0.4770 | 0.4791 | 0.4813 | 0.4834 | 0.4856 |
| 26    | sin      | 0.4384 | 0.4399 | 0.4415 | 0.4431 | 0.4446 | 0.4462 | 0.4478 | 0.4493 | 0.4509 | 0.4524 |
|       | cos      | 0.8958 | 0.8950 | 0.8943 | 0.8935 | 0.8927 | 0.8919 | 0.8912 | 0.8904 | 0.8896 | 0.8888 |
|       | tan      | 0.4877 | 0.4899 | 0.4921 | 0.4942 | 0.4964 | 0.4986 | 0.5008 | 0.5029 | 0.5051 | 0.5073 |
| 27    | sin      | 0.4540 | 0.4555 | 0.4571 | 0.4586 | 0.4602 | 0.4617 | 0.4633 | 0.4648 | 0.4664 | 0.4679 |
|       | cos      | 0.8910 | 0.8902 | 0.8894 | 0.8886 | 0.8878 | 0.8870 | 0.8862 | 0.8854 | 0.8846 | 0.8838 |
|       | tan      | 0.5095 | 0.5117 | 0.5139 | 0.5161 | 0.5184 | 0.5206 | 0.5228 | 0.5250 | 0.5272 | 0.5295 |
| 28    | sin      | 0.4695 | 0.4710 | 0.4726 | 0.4741 | 0.4756 | 0.4772 | 0.4787 | 0.4802 | 0.4818 | 0.4833 |
|       | cos      | 0.8829 | 0.8821 | 0.8813 | 0.8805 | 0.8796 | 0.8788 | 0.8780 | 0.8771 | 0.8763 | 0.8755 |
|       | tan      | 0.5317 | 0.5340 | 0.5362 | 0.5384 | 0.5407 | 0.5430 | 0.5452 | 0.5475 | 0.5498 | 0.5520 |
| 29    | sin      | 0.4848 | 0.4863 | 0.4879 | 0.4894 | 0.4909 | 0.4924 | 0.4939 | 0.4955 | 0.4970 | 0.4985 |
|       | cos      | 0.8746 | 0.8738 | 0.8729 | 0.8721 | 0.8712 | 0.8704 | 0.8695 | 0.8686 | 0.8678 | 0.8669 |
|       | tan      | 0.5543 | 0.5566 | 0.5589 | 0.5612 | 0.5635 | 0.5658 | 0.5681 | 0.5704 | 0.5727 | 0.5750 |
| Degs. | Function | 0'     | 6'     | 12'    | 18'    | 24'    | 30'    | 36'    | 42'    | 48'    | 54'    |

# Natural Sines, Cosines and Tangents

30°-44.9°

| Degs. | Function | 0.0°   | 0.1°   | 0.2°   | 0.3°   | 0.4°   | 0.5°   | 0.6°   | 0.7°   | 0.8°   | 0.9°   |
|-------|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 30    | sin      | 0.5000 | 0.5015 | 0.5030 | 0.5045 | 0.5060 | 0.5075 | 0.5090 | 0.5105 | 0.5120 | 0.5135 |
|       | cos      | 0.8660 | 0.8652 | 0.8643 | 0.8634 | 0.8625 | 0.8616 | 0.8607 | 0.8599 | 0.8590 | 0.8581 |
|       | tan      | 0.5774 | 0.5797 | 0.5820 | 0.5844 | 0.5867 | 0.5890 | 0.5914 | 0.5938 | 0.5961 | 0.5985 |
| 31    | sin      | 0.5150 | 0.5165 | 0.5180 | 0.5195 | 0.5210 | 0.5225 | 0.5240 | 0.5255 | 0.5270 | 0.5284 |
|       | cos      | 0.8572 | 0.8563 | 0.8554 | 0.8545 | 0.8536 | 0.8526 | 0.8517 | 0.8508 | 0.8499 | 0.8490 |
|       | tan      | 0.6009 | 0.6032 | 0.6056 | 0.6080 | 0.6104 | 0.6128 | 0.6152 | 0.6176 | 0.6200 | 0.6224 |
| 32    | sin      | 0.5299 | 0.5314 | 0.5329 | 0.5344 | 0.5358 | 0.5373 | 0.5388 | 0.5402 | 0.5417 | 0.5432 |
|       | cos      | 0.8480 | 0.8471 | 0.8462 | 0.8453 | 0.8443 | 0.8434 | 0.8425 | 0.8415 | 0.8406 | 0.8396 |
|       | tan      | 0.6249 | 0.6273 | 0.6297 | 0.6322 | 0.6346 | 0.6371 | 0.6395 | 0.6420 | 0.6445 | 0.6469 |
| 33    | sin      | 0.5446 | 0.5461 | 0.5476 | 0.5490 | 0.5505 | 0.5519 | 0.5534 | 0.5548 | 0.5563 | 0.5577 |
|       | cos      | 0.8387 | 0.8377 | 0.8368 | 0.8358 | 0.8348 | 0.8339 | 0.8329 | 0.8320 | 0.8310 | 0.8300 |
|       | tan      | 0.6494 | 0.6519 | 0.6544 | 0.6569 | 0.6594 | 0.6619 | 0.6644 | 0.6669 | 0.6694 | 0.6720 |
| 34    | sin      | 0.5592 | 0.5606 | 0.5621 | 0.5635 | 0.5650 | 0.5664 | 0.5678 | 0.5693 | 0.5707 | 0.5721 |
|       | cos      | 0.8290 | 0.8281 | 0.8271 | 0.8261 | 0.8251 | 0.8241 | 0.8231 | 0.8221 | 0.8211 | 0.8202 |
|       | tan      | 0.6745 | 0.6771 | 0.6796 | 0.6822 | 0.6847 | 0.6873 | 0.6899 | 0.6924 | 0.6950 | 0.6976 |
| 35    | sin      | 0.5736 | 0.5750 | 0.5764 | 0.5779 | 0.5793 | 0.5807 | 0.5821 | 0.5835 | 0.5850 | 0.5864 |
|       | cos      | 0.8192 | 0.8181 | 0.8171 | 0.8161 | 0.8151 | 0.8141 | 0.8131 | 0.8121 | 0.8111 | 0.8100 |
|       | tan      | 0.7002 | 0.7028 | 0.7054 | 0.7080 | 0.7107 | 0.7133 | 0.7159 | 0.7186 | 0.7212 | 0.7239 |
| 36    | sin      | 0.5878 | 0.5892 | 0.5906 | 0.5920 | 0.5934 | 0.5948 | 0.5962 | 0.5976 | 0.5990 | 0.6004 |
|       | cos      | 0.8090 | 0.8080 | 0.8070 | 0.8059 | 0.8049 | 0.8039 | 0.8028 | 0.8018 | 0.8007 | 0.7997 |
|       | tan      | 0.7265 | 0.7292 | 0.7319 | 0.7346 | 0.7373 | 0.7400 | 0.7427 | 0.7454 | 0.7481 | 0.7508 |
| 37    | sin      | 0.6018 | 0.6032 | 0.6046 | 0.6060 | 0.6074 | 0.6088 | 0.6101 | 0.6115 | 0.6129 | 0.6143 |
|       | cos      | 0.7986 | 0.7976 | 0.7965 | 0.7955 | 0.7944 | 0.7934 | 0.7923 | 0.7912 | 0.7902 | 0.7891 |
|       | tan      | 0.7536 | 0.7563 | 0.7590 | 0.7618 | 0.7646 | 0.7673 | 0.7701 | 0.7729 | 0.7757 | 0.7785 |
| 38    | sin      | 0.6157 | 0.6170 | 0.6184 | 0.6198 | 0.6211 | 0.6225 | 0.6239 | 0.6252 | 0.6266 | 0.6280 |
|       | cos      | 0.7880 | 0.7869 | 0.7859 | 0.7848 | 0.7837 | 0.7826 | 0.7815 | 0.7804 | 0.7793 | 0.7782 |
|       | tan      | 0.7813 | 0.7841 | 0.7869 | 0.7898 | 0.7926 | 0.7954 | 0.7983 | 0.8012 | 0.8040 | 0.8069 |
| 39    | sin      | 0.6293 | 0.6307 | 0.6320 | 0.6334 | 0.6347 | 0.6361 | 0.6374 | 0.6388 | 0.6401 | 0.6414 |
|       | cos      | 0.7771 | 0.7760 | 0.7749 | 0.7738 | 0.7727 | 0.7716 | 0.7705 | 0.7694 | 0.7683 | 0.7672 |
|       | tan      | 0.8098 | 0.8127 | 0.8156 | 0.8185 | 0.8214 | 0.8243 | 0.8273 | 0.8302 | 0.8332 | 0.8361 |
| 40    | sin      | 0.6428 | 0.6441 | 0.6455 | 0.6468 | 0.6481 | 0.6494 | 0.6508 | 0.6521 | 0.6534 | 0.6547 |
|       | cos      | 0.7660 | 0.7649 | 0.7638 | 0.7627 | 0.7615 | 0.7604 | 0.7593 | 0.7581 | 0.7570 | 0.7559 |
|       | tan      | 0.8391 | 0.8421 | 0.8451 | 0.8481 | 0.8511 | 0.8541 | 0.8571 | 0.8601 | 0.8632 | 0.8662 |
| 41    | sin      | 0.6561 | 0.6574 | 0.6587 | 0.6600 | 0.6613 | 0.6626 | 0.6639 | 0.6652 | 0.6665 | 0.6678 |
|       | cos      | 0.7547 | 0.7536 | 0.7524 | 0.7513 | 0.7501 | 0.7490 | 0.7478 | 0.7466 | 0.7455 | 0.7443 |
|       | tan      | 0.8693 | 0.8724 | 0.8754 | 0.8785 | 0.8816 | 0.8847 | 0.8878 | 0.8910 | 0.8941 | 0.8972 |
| 42    | sin      | 0.6691 | 0.6704 | 0.6717 | 0.6730 | 0.6743 | 0.6756 | 0.6769 | 0.6782 | 0.6794 | 0.6807 |
|       | cos      | 0.7431 | 0.7420 | 0.7408 | 0.7396 | 0.7385 | 0.7373 | 0.7361 | 0.7349 | 0.7337 | 0.7325 |
|       | tan      | 0.9004 | 0.9036 | 0.9067 | 0.9099 | 0.9131 | 0.9163 | 0.9195 | 0.9228 | 0.9260 | 0.9293 |
| 43    | sin      | 0.6820 | 0.6833 | 0.6845 | 0.6858 | 0.6871 | 0.6884 | 0.6896 | 0.6909 | 0.6921 | 0.6934 |
|       | cos      | 0.7314 | 0.7302 | 0.7290 | 0.7278 | 0.7266 | 0.7254 | 0.7242 | 0.7230 | 0.7218 | 0.7206 |
|       | tan      | 0.9325 | 0.9358 | 0.9391 | 0.9424 | 0.9457 | 0.9490 | 0.9523 | 0.9556 | 0.9590 | 0.9623 |
| 44    | sin      | 0.6947 | 0.6959 | 0.6972 | 0.6984 | 0.6997 | 0.7009 | 0.7022 | 0.7034 | 0.7046 | 0.7059 |
|       | cos      | 0.7193 | 0.7181 | 0.7169 | 0.7157 | 0.7145 | 0.7133 | 0.7120 | 0.7108 | 0.7096 | 0.7083 |
|       | tan      | 0.9657 | 0.9691 | 0.9725 | 0.9759 | 0.9793 | 0.9827 | 0.9861 | 0.9896 | 0.9930 | 0.9965 |
| Degs. | Function | 0'     | 6'     | 12'    | 18'    | 24'    | 30'    | 36'    | 42'    | 48'    | 54'    |

# Natural Sines, Cosines and Tangents

45°-59.9°

| Degs. | Function | 0.0°   | 0.1°   | 0.2°   | 0.3°   | 0.4°   | 0.5°   | 0.6°   | 0.7°   | 0.8°   | 0.9°   |
|-------|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 45    | sin      | 0.7071 | 0.7083 | 0.7096 | 0.7108 | 0.7120 | 0.7133 | 0.7145 | 0.7157 | 0.7169 | 0.7181 |
|       | cos      | 0.7071 | 0.7059 | 0.7046 | 0.7034 | 0.7022 | 0.7009 | 0.6997 | 0.6984 | 0.6972 | 0.6959 |
|       | tan      | 1.0000 | 1.0035 | 1.0070 | 1.0105 | 1.0141 | 1.0176 | 1.0212 | 1.0247 | 1.0283 | 1.0319 |
| 46    | sin      | 0.7193 | 0.7206 | 0.7218 | 0.7230 | 0.7242 | 0.7254 | 0.7266 | 0.7278 | 0.7290 | 0.7302 |
|       | cos      | 0.6947 | 0.6934 | 0.6921 | 0.6909 | 0.6896 | 0.6884 | 0.6871 | 0.6858 | 0.6845 | 0.6833 |
|       | tan      | 1.0355 | 1.0392 | 1.0428 | 1.0464 | 1.0501 | 1.0538 | 1.0575 | 1.0612 | 1.0649 | 1.0686 |
| 47    | sin      | 0.7314 | 0.7325 | 0.7337 | 0.7349 | 0.7361 | 0.7373 | 0.7385 | 0.7396 | 0.7408 | 0.7420 |
|       | cos      | 0.6820 | 0.6807 | 0.6794 | 0.6782 | 0.6769 | 0.6756 | 0.6743 | 0.6730 | 0.6717 | 0.6704 |
|       | tan      | 1.0724 | 1.0761 | 1.0799 | 1.0837 | 1.0875 | 1.0913 | 1.0951 | 1.0990 | 1.1028 | 1.1067 |
| 48    | sin      | 0.7431 | 0.7443 | 0.7455 | 0.7466 | 0.7478 | 0.7490 | 0.7501 | 0.7513 | 0.7524 | 0.7536 |
|       | cos      | 0.6691 | 0.6678 | 0.6665 | 0.6652 | 0.6639 | 0.6626 | 0.6613 | 0.6600 | 0.6587 | 0.6574 |
|       | tan      | 1.1106 | 1.1145 | 1.1184 | 1.1224 | 1.1263 | 1.1303 | 1.1343 | 1.1383 | 1.1423 | 1.1463 |
| 49    | sin      | 0.7547 | 0.7559 | 0.7570 | 0.7581 | 0.7593 | 0.7604 | 0.7615 | 0.7627 | 0.7638 | 0.7649 |
|       | cos      | 0.6561 | 0.6547 | 0.6534 | 0.6521 | 0.6508 | 0.6494 | 0.6481 | 0.6468 | 0.6455 | 0.6441 |
|       | tan      | 1.1504 | 1.1544 | 1.1585 | 1.1626 | 1.1667 | 1.1708 | 1.1750 | 1.1792 | 1.1833 | 1.1875 |
| 50    | sin      | 0.7660 | 0.7672 | 0.7683 | 0.7694 | 0.7705 | 0.7716 | 0.7727 | 0.7738 | 0.7749 | 0.7760 |
|       | cos      | 0.6428 | 0.6414 | 0.6401 | 0.6388 | 0.6374 | 0.6361 | 0.6347 | 0.6334 | 0.6320 | 0.6307 |
|       | tan      | 1.1918 | 1.1960 | 1.2002 | 1.2045 | 1.2088 | 1.2131 | 1.2174 | 1.2218 | 1.2261 | 1.2305 |
| 51    | sin      | 0.7771 | 0.7782 | 0.7793 | 0.7804 | 0.7815 | 0.7826 | 0.7837 | 0.7848 | 0.7859 | 0.7869 |
|       | cos      | 0.6293 | 0.6280 | 0.6266 | 0.6252 | 0.6239 | 0.6225 | 0.6211 | 0.6198 | 0.6184 | 0.6170 |
|       | tan      | 1.2349 | 1.2393 | 1.2437 | 1.2482 | 1.2527 | 1.2572 | 1.2617 | 1.2662 | 1.2708 | 1.2753 |
| 52    | sin      | 0.7880 | 0.7891 | 0.7902 | 0.7912 | 0.7923 | 0.7934 | 0.7944 | 0.7955 | 0.7965 | 0.7976 |
|       | cos      | 0.6157 | 0.6143 | 0.6129 | 0.6115 | 0.6101 | 0.6088 | 0.6074 | 0.6060 | 0.6046 | 0.6032 |
|       | tan      | 1.2799 | 1.2846 | 1.2892 | 1.2938 | 1.2985 | 1.3032 | 1.3079 | 1.3127 | 1.3175 | 1.3222 |
| 53    | sin      | 0.7986 | 0.7997 | 0.8007 | 0.8018 | 0.8028 | 0.8039 | 0.8049 | 0.8059 | 0.8070 | 0.8080 |
|       | cos      | 0.6018 | 0.6004 | 0.5990 | 0.5976 | 0.5962 | 0.5948 | 0.5934 | 0.5920 | 0.5906 | 0.5892 |
|       | tan      | 1.3270 | 1.3319 | 1.3367 | 1.3416 | 1.3465 | 1.3514 | 1.3564 | 1.3613 | 1.3663 | 1.3713 |
| 54    | sin      | 0.8090 | 0.8100 | 0.8111 | 0.8121 | 0.8131 | 0.8141 | 0.8151 | 0.8161 | 0.8171 | 0.8181 |
|       | cos      | 0.5878 | 0.5864 | 0.5850 | 0.5835 | 0.5821 | 0.5807 | 0.5793 | 0.5779 | 0.5764 | 0.5750 |
|       | tan      | 1.3764 | 1.3814 | 1.3865 | 1.3916 | 1.3968 | 1.4019 | 1.4071 | 1.4124 | 1.4176 | 1.4229 |
| 55    | sin      | 0.8192 | 0.8202 | 0.8211 | 0.8221 | 0.8231 | 0.8241 | 0.8251 | 0.8261 | 0.8271 | 0.8281 |
|       | cos      | 0.5736 | 0.5721 | 0.5707 | 0.5693 | 0.5678 | 0.5664 | 0.5650 | 0.5635 | 0.5621 | 0.5606 |
|       | tan      | 1.4281 | 1.4335 | 1.4388 | 1.4442 | 1.4496 | 1.4550 | 1.4605 | 1.4659 | 1.4715 | 1.4770 |
| 56    | sin      | 0.8290 | 0.8300 | 0.8310 | 0.8320 | 0.8329 | 0.8339 | 0.8348 | 0.8358 | 0.8368 | 0.8377 |
|       | cos      | 0.5592 | 0.5577 | 0.5563 | 0.5548 | 0.5534 | 0.5519 | 0.5505 | 0.5490 | 0.5476 | 0.5461 |
|       | tan      | 1.4826 | 1.4882 | 1.4938 | 1.4994 | 1.5051 | 1.5108 | 1.5166 | 1.5224 | 1.5282 | 1.5340 |
| 57    | sin      | 0.8387 | 0.8396 | 0.8406 | 0.8415 | 0.8425 | 0.8434 | 0.8443 | 0.8453 | 0.8462 | 0.8471 |
|       | cos      | 0.5446 | 0.5432 | 0.5417 | 0.5402 | 0.5388 | 0.5373 | 0.5358 | 0.5344 | 0.5329 | 0.5314 |
|       | tan      | 1.5399 | 1.5458 | 1.5517 | 1.5577 | 1.5637 | 1.5697 | 1.5757 | 1.5818 | 1.5880 | 1.5941 |
| 58    | sin      | 0.8480 | 0.8490 | 0.8499 | 0.8508 | 0.8517 | 0.8526 | 0.8536 | 0.8545 | 0.8554 | 0.8563 |
|       | cos      | 0.5299 | 0.5284 | 0.5270 | 0.5255 | 0.5240 | 0.5225 | 0.5210 | 0.5195 | 0.5180 | 0.5165 |
|       | tan      | 1.6003 | 1.6066 | 1.6128 | 1.6191 | 1.6255 | 1.6319 | 1.6383 | 1.6447 | 1.6512 | 1.6577 |
| 59    | sin      | 0.8572 | 0.8581 | 0.8590 | 0.8599 | 0.8607 | 0.8616 | 0.8625 | 0.8634 | 0.8643 | 0.8652 |
|       | cos      | 0.5150 | 0.5135 | 0.5120 | 0.5105 | 0.5090 | 0.5075 | 0.5060 | 0.5045 | 0.5030 | 0.5015 |
|       | tan      | 1.6643 | 1.6709 | 1.6775 | 1.6842 | 1.6909 | 1.6977 | 1.7045 | 1.7113 | 1.7182 | 1.7251 |
| Degs. | Function | 0'     | 6'     | 12'    | 18'    | 24'    | 30'    | 36'    | 42'    | 48'    | 54'    |

# Natural Sines, Cosines and Tangents

60°-74.9°

| Degs. | Function | 0.0°   | 0.1°   | 0.2°   | 0.3°   | 0.4°   | 0.5°   | 0.6°   | 0.7°   | 0.8°   | 0.9°   |
|-------|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 60    | sin      | 0.8660 | 0.8669 | 0.8678 | 0.8686 | 0.8695 | 0.8704 | 0.8712 | 0.8721 | 0.8729 | 0.8738 |
|       | cos      | 0.5000 | 0.4985 | 0.4970 | 0.4955 | 0.4939 | 0.4924 | 0.4909 | 0.4894 | 0.4879 | 0.4863 |
|       | tan      | 1.7321 | 1.7391 | 1.7461 | 1.7532 | 1.7603 | 1.7675 | 1.7747 | 1.7820 | 1.7893 | 1.7966 |
| 61    | sin      | 0.8746 | 0.8755 | 0.8763 | 0.8771 | 0.8780 | 0.8788 | 0.8796 | 0.8805 | 0.8813 | 0.8821 |
|       | cos      | 0.4848 | 0.4833 | 0.4818 | 0.4802 | 0.4787 | 0.4772 | 0.4756 | 0.4741 | 0.4726 | 0.4710 |
|       | tan      | 1.8040 | 1.8115 | 1.8190 | 1.8265 | 1.8341 | 1.8418 | 1.8495 | 1.8572 | 1.8650 | 1.8728 |
| 62    | sin      | 0.8829 | 0.8838 | 0.8846 | 0.8854 | 0.8862 | 0.8870 | 0.8878 | 0.8886 | 0.8894 | 0.8902 |
|       | cos      | 0.4695 | 0.4679 | 0.4664 | 0.4648 | 0.4633 | 0.4617 | 0.4602 | 0.4586 | 0.4571 | 0.4555 |
|       | tan      | 1.8807 | 1.8887 | 1.8967 | 1.9047 | 1.9128 | 1.9210 | 1.9292 | 1.9375 | 1.9458 | 1.9542 |
| 63    | sin      | 0.8910 | 0.8918 | 0.8926 | 0.8934 | 0.8942 | 0.8949 | 0.8957 | 0.8965 | 0.8973 | 0.8980 |
|       | cos      | 0.4540 | 0.4524 | 0.4509 | 0.4493 | 0.4478 | 0.4462 | 0.4446 | 0.4431 | 0.4415 | 0.4399 |
|       | tan      | 1.9626 | 1.9711 | 1.9797 | 1.9883 | 1.9970 | 2.0057 | 2.0145 | 2.0233 | 2.0323 | 2.0413 |
| 64    | sin      | 0.8988 | 0.8996 | 0.9003 | 0.9011 | 0.9018 | 0.9026 | 0.9033 | 0.9041 | 0.9048 | 0.9056 |
|       | cos      | 0.4384 | 0.4368 | 0.4352 | 0.4337 | 0.4321 | 0.4305 | 0.4289 | 0.4274 | 0.4258 | 0.4242 |
|       | tan      | 2.0503 | 2.0594 | 2.0686 | 2.0778 | 2.0872 | 2.0965 | 2.1060 | 2.1155 | 2.1251 | 2.1348 |
| 65    | sin      | 0.9063 | 0.9070 | 0.9078 | 0.9085 | 0.9092 | 0.9100 | 0.9107 | 0.9114 | 0.9121 | 0.9128 |
|       | cos      | 0.4226 | 0.4210 | 0.4195 | 0.4179 | 0.4163 | 0.4147 | 0.4131 | 0.4115 | 0.4099 | 0.4083 |
|       | tan      | 2.1445 | 2.1543 | 2.1642 | 2.1742 | 2.1842 | 2.1943 | 2.2045 | 2.2148 | 2.2251 | 2.2355 |
| 66    | sin      | 0.9135 | 0.9143 | 0.9150 | 0.9157 | 0.9164 | 0.9171 | 0.9178 | 0.9184 | 0.9191 | 0.9198 |
|       | cos      | 0.4067 | 0.4051 | 0.4035 | 0.4019 | 0.4003 | 0.3987 | 0.3971 | 0.3955 | 0.3939 | 0.3923 |
|       | tan      | 2.2460 | 2.2566 | 2.2673 | 2.2781 | 2.2889 | 2.2998 | 2.3109 | 2.3220 | 2.3332 | 2.3445 |
| 67    | sin      | 0.9205 | 0.9212 | 0.9219 | 0.9225 | 0.9232 | 0.9239 | 0.9245 | 0.9252 | 0.9259 | 0.9265 |
|       | cos      | 0.3907 | 0.3891 | 0.3875 | 0.3859 | 0.3843 | 0.3827 | 0.3811 | 0.3795 | 0.3778 | 0.3762 |
|       | tan      | 2.3559 | 2.3673 | 2.3789 | 2.3906 | 2.4023 | 2.4142 | 2.4262 | 2.4383 | 2.4504 | 2.4627 |
| 68    | sin      | 0.9272 | 0.9278 | 0.9285 | 0.9291 | 0.9298 | 0.9304 | 0.9311 | 0.9317 | 0.9323 | 0.9330 |
|       | cos      | 0.3746 | 0.3730 | 0.3714 | 0.3697 | 0.3681 | 0.3665 | 0.3649 | 0.3633 | 0.3616 | 0.3600 |
|       | tan      | 2.4751 | 2.4876 | 2.5002 | 2.5129 | 2.5257 | 2.5386 | 2.5517 | 2.5649 | 2.5782 | 2.5916 |
| 69    | sin      | 0.9336 | 0.9342 | 0.9348 | 0.9354 | 0.9361 | 0.9367 | 0.9373 | 0.9379 | 0.9385 | 0.9391 |
|       | cos      | 0.3584 | 0.3567 | 0.3551 | 0.3535 | 0.3518 | 0.3502 | 0.3486 | 0.3469 | 0.3453 | 0.3437 |
|       | tan      | 2.6051 | 2.6187 | 2.6325 | 2.6464 | 2.6605 | 2.6746 | 2.6889 | 2.7034 | 2.7179 | 2.7326 |
| 70    | sin      | 0.9397 | 0.9403 | 0.9409 | 0.9415 | 0.9421 | 0.9426 | 0.9432 | 0.9438 | 0.9444 | 0.9449 |
|       | cos      | 0.3420 | 0.3404 | 0.3387 | 0.3371 | 0.3355 | 0.3338 | 0.3322 | 0.3305 | 0.3289 | 0.3272 |
|       | tan      | 2.7475 | 2.7625 | 2.7776 | 2.7929 | 2.8083 | 2.8239 | 2.8397 | 2.8556 | 2.8716 | 2.8878 |
| 71    | sin      | 0.9455 | 0.9461 | 0.9466 | 0.9472 | 0.9478 | 0.9483 | 0.9489 | 0.9494 | 0.9500 | 0.9505 |
|       | cos      | 0.3256 | 0.3239 | 0.3223 | 0.3206 | 0.3190 | 0.3173 | 0.3156 | 0.3140 | 0.3123 | 0.3107 |
|       | tan      | 2.9042 | 2.9208 | 2.9375 | 2.9544 | 2.9714 | 2.9887 | 3.0061 | 3.0237 | 3.0415 | 3.0595 |
| 72    | sin      | 0.9511 | 0.9516 | 0.9521 | 0.9527 | 0.9532 | 0.9537 | 0.9542 | 0.9548 | 0.9553 | 0.9558 |
|       | cos      | 0.3090 | 0.3074 | 0.3057 | 0.3040 | 0.3024 | 0.3007 | 0.2990 | 0.2974 | 0.2957 | 0.2940 |
|       | tan      | 3.0777 | 3.0961 | 3.1146 | 3.1334 | 3.1524 | 3.1716 | 3.1910 | 3.2106 | 3.2305 | 3.2506 |
| 73    | sin      | 0.9563 | 0.9568 | 0.9573 | 0.9578 | 0.9583 | 0.9588 | 0.9593 | 0.9598 | 0.9603 | 0.9608 |
|       | cos      | 0.2924 | 0.2907 | 0.2890 | 0.2874 | 0.2857 | 0.2840 | 0.2823 | 0.2807 | 0.2790 | 0.2773 |
|       | tan      | 3.2709 | 3.2914 | 3.3122 | 3.3332 | 3.3544 | 3.3759 | 3.3977 | 3.4197 | 3.4420 | 3.4646 |
| 74    | sin      | 0.9613 | 0.9617 | 0.9622 | 0.9627 | 0.9632 | 0.9636 | 0.9641 | 0.9646 | 0.9650 | 0.9655 |
|       | cos      | 0.2756 | 0.2740 | 0.2723 | 0.2706 | 0.2689 | 0.2672 | 0.2656 | 0.2639 | 0.2622 | 0.2605 |
|       | tan      | 3.4874 | 3.5105 | 3.5339 | 3.5576 | 3.5816 | 3.6059 | 3.6305 | 3.6554 | 3.6806 | 3.7062 |
| Degs. | Function | 0'     | 6'     | 12'    | 18'    | 24'    | 30'    | 36'    | 42'    | 48'    | 54'    |

# Natural Sines, Cosines and Tangents

75°-89.9°

| Degs. | Function | 0.0°   | 0.1°   | 0.2°   | 0.3°   | 0.4°   | 0.5°   | 0.6°   | 0.7°   | 0.8°   | 0.9°   |
|-------|----------|--------|--------|--------|--------|--------|--------|--------|--------|--------|--------|
| 75    | sin      | 0.9659 | 0.9664 | 0.9668 | 0.9673 | 0.9677 | 0.9681 | 0.9686 | 0.9690 | 0.9694 | 0.9699 |
|       | cos      | 0.2588 | 0.2571 | 0.2554 | 0.2538 | 0.2521 | 0.2504 | 0.2487 | 0.2470 | 0.2453 | 0.2436 |
|       | tan      | 3.7321 | 3.7583 | 3.7848 | 3.8118 | 3.8391 | 3.8667 | 3.8947 | 3.9232 | 3.9520 | 3.9812 |
| 76    | sin      | 0.9703 | 0.9707 | 0.9711 | 0.9715 | 0.9720 | 0.9724 | 0.9728 | 0.9732 | 0.9736 | 0.9740 |
|       | cos      | 0.2419 | 0.2402 | 0.2385 | 0.2368 | 0.2351 | 0.2334 | 0.2317 | 0.2300 | 0.2284 | 0.2267 |
|       | tan      | 4.0108 | 4.0408 | 4.0713 | 4.1022 | 4.1335 | 4.1653 | 4.1976 | 4.2303 | 4.2635 | 4.2972 |
| 77    | sin      | 0.9744 | 0.9748 | 0.9751 | 0.9755 | 0.9759 | 0.9763 | 0.9767 | 0.9770 | 0.9774 | 0.9778 |
|       | cos      | 0.2250 | 0.2232 | 0.2215 | 0.2198 | 0.2181 | 0.2164 | 0.2147 | 0.2130 | 0.2113 | 0.2096 |
|       | tan      | 4.3315 | 4.3662 | 4.4015 | 4.4374 | 4.4737 | 4.5107 | 4.5483 | 4.5864 | 4.6252 | 4.6646 |
| 78    | sin      | 0.9781 | 0.9785 | 0.9789 | 0.9792 | 0.9796 | 0.9799 | 0.9803 | 0.9806 | 0.9810 | 0.9813 |
|       | cos      | 0.2079 | 0.2062 | 0.2045 | 0.2028 | 0.2011 | 0.1994 | 0.1977 | 0.1959 | 0.1942 | 0.1925 |
|       | tan      | 4.7046 | 4.7453 | 4.7867 | 4.8288 | 4.8716 | 4.9152 | 4.9594 | 5.0045 | 5.0504 | 5.0970 |
| 79    | sin      | 0.9816 | 0.9820 | 0.9823 | 0.9826 | 0.9829 | 0.9833 | 0.9836 | 0.9839 | 0.9842 | 0.9845 |
|       | cos      | 0.1908 | 0.1891 | 0.1874 | 0.1857 | 0.1840 | 0.1822 | 0.1805 | 0.1788 | 0.1771 | 0.1754 |
|       | tan      | 5.1446 | 5.1929 | 5.2422 | 5.2924 | 5.3435 | 5.3955 | 5.4486 | 5.5026 | 5.5578 | 5.6140 |
| 80    | sin      | 0.9848 | 0.9851 | 0.9854 | 0.9857 | 0.9860 | 0.9863 | 0.9866 | 0.9869 | 0.9871 | 0.9874 |
|       | cos      | 0.1736 | 0.1719 | 0.1702 | 0.1685 | 0.1668 | 0.1650 | 0.1633 | 0.1616 | 0.1599 | 0.1582 |
|       | tan      | 5.6713 | 5.7297 | 5.7894 | 5.8502 | 5.9124 | 5.9758 | 6.0405 | 6.1066 | 6.1742 | 6.2432 |
| 81    | sin      | 0.9877 | 0.9880 | 0.9882 | 0.9885 | 0.9888 | 0.9890 | 0.9893 | 0.9895 | 0.9898 | 0.9900 |
|       | cos      | 0.1561 | 0.1547 | 0.1530 | 0.1513 | 0.1495 | 0.1478 | 0.1461 | 0.1444 | 0.1426 | 0.1409 |
|       | tan      | 6.3138 | 6.3859 | 6.4596 | 6.5350 | 6.6122 | 6.6912 | 6.7720 | 6.8548 | 6.9395 | 7.0264 |
| 82    | sin      | 0.9903 | 0.9905 | 0.9907 | 0.9910 | 0.9912 | 0.9914 | 0.9917 | 0.9919 | 0.9921 | 0.9923 |
|       | cos      | 0.1392 | 0.1374 | 0.1357 | 0.1340 | 0.1323 | 0.1305 | 0.1288 | 0.1271 | 0.1253 | 0.1236 |
|       | tan      | 7.1154 | 7.2066 | 7.3002 | 7.3962 | 7.4947 | 7.5958 | 7.6996 | 7.8062 | 7.9158 | 8.0285 |
| 83    | sin      | 0.9925 | 0.9928 | 0.9930 | 0.9932 | 0.9934 | 0.9936 | 0.9938 | 0.9940 | 0.9942 | 0.9943 |
|       | cos      | 0.1219 | 0.1201 | 0.1184 | 0.1167 | 0.1149 | 0.1132 | 0.1115 | 0.1097 | 0.1080 | 0.1063 |
|       | tan      | 8.1443 | 8.2636 | 8.3863 | 8.5126 | 8.6427 | 8.7769 | 8.9152 | 9.0579 | 9.2052 | 9.3572 |
| 84    | sin      | 0.9945 | 0.9947 | 0.9949 | 0.9951 | 0.9952 | 0.9954 | 0.9956 | 0.9957 | 0.9959 | 0.9960 |
|       | cos      | 0.1045 | 0.1028 | 0.1011 | 0.0993 | 0.0976 | 0.0958 | 0.0941 | 0.0924 | 0.0906 | 0.0889 |
|       | tan      | 9.5144 | 9.6768 | 9.8448 | 10.02  | 10.20  | 10.39  | 10.58  | 10.78  | 10.99  | 11.20  |
| 85    | sin      | 0.9962 | 0.9963 | 0.9965 | 0.9966 | 0.9968 | 0.9969 | 0.9971 | 0.9972 | 0.9973 | 0.9974 |
|       | cos      | 0.0872 | 0.0854 | 0.0837 | 0.0819 | 0.0802 | 0.0785 | 0.0767 | 0.0750 | 0.0732 | 0.0715 |
|       | tan      | 11.43  | 11.66  | 11.91  | 12.16  | 12.43  | 12.71  | 13.00  | 13.30  | 13.62  | 13.95  |
| 86    | sin      | 0.9976 | 0.9977 | 0.9978 | 0.9979 | 0.9980 | 0.9981 | 0.9982 | 0.9983 | 0.9984 | 0.9985 |
|       | cos      | 0.0698 | 0.0680 | 0.0663 | 0.0645 | 0.0628 | 0.0610 | 0.0593 | 0.0576 | 0.0558 | 0.0541 |
|       | tan      | 14.30  | 14.67  | 15.06  | 15.46  | 15.89  | 16.35  | 16.83  | 17.34  | 17.89  | 18.46  |
| 87    | sin      | 0.9986 | 0.9987 | 0.9988 | 0.9989 | 0.9990 | 0.9990 | 0.9991 | 0.9992 | 0.9993 | 0.9993 |
|       | cos      | 0.0523 | 0.0506 | 0.0488 | 0.0471 | 0.0454 | 0.0436 | 0.0419 | 0.0401 | 0.0384 | 0.0366 |
|       | tan      | 19.08  | 19.74  | 20.45  | 21.20  | 22.02  | 22.90  | 23.86  | 24.90  | 26.03  | 27.27  |
| 88    | sin      | 0.9994 | 0.9995 | 0.9995 | 0.9996 | 0.9996 | 0.9997 | 0.9997 | 0.9997 | 0.9998 | 0.9998 |
|       | cos      | 0.0349 | 0.0332 | 0.0314 | 0.0297 | 0.0279 | 0.0262 | 0.0244 | 0.0227 | 0.0209 | 0.0192 |
|       | tan      | 28.64  | 30.14  | 31.82  | 33.69  | 35.80  | 38.19  | 40.92  | 44.07  | 47.74  | 52.08  |
| 89    | sin      | 0.9998 | 0.9999 | 0.9999 | 0.9999 | 0.9999 | 1.000  | 1.000  | 1.000  | 1.000  | 1.000  |
|       | cos      | 0.0175 | 0.0157 | 0.0140 | 0.0122 | 0.0105 | 0.0087 | 0.0070 | 0.0052 | 0.0035 | 0.0017 |
|       | tan      | 57.29  | 63.66  | 71.62  | 81.85  | 95.49  | 114.6  | 143.2  | 191.0  | 286.5  | 573.0  |
| Degs. | Function | 0'     | 6'     | 12'    | 18'    | 24'    | 30'    | 36'    | 42'    | 48'    | 54'    |



## Squares, Cubes and Roots

Roots of numbers other than those given directly may be found by the following relations:

$$\begin{aligned} \sqrt{100n} &= 10 \sqrt{n}; & \sqrt{1000n} &= 10 \sqrt{10n}; & \sqrt{\frac{1}{10}n} &= \frac{1}{10} \sqrt{10n}; & \sqrt{\frac{1}{100}n} &= \frac{1}{10} \sqrt{n}, \\ \sqrt[3]{\frac{1}{1000}n} &= \frac{1}{100} \sqrt[3]{10n}; & \sqrt[3]{1000n} &= 10 \sqrt[3]{n}; & \sqrt[3]{10,000n} &= 10 \sqrt[3]{10n}; & \sqrt[3]{100,000n} &= 10 \sqrt[3]{100n}; \\ 10 \sqrt[3]{100n} &= \sqrt[3]{\frac{1}{10}n} = \frac{1}{10} \sqrt[3]{100n}; & \sqrt[3]{\frac{1}{10}n} &= \frac{1}{10} \sqrt[3]{10n}; & \sqrt[3]{\frac{1}{100}n} &= \frac{1}{10} \sqrt[3]{n}. \end{aligned}$$

| <i>n</i> | <i>n</i> <sup>2</sup> | $\sqrt{n}$ | $\sqrt{10n}$ | <i>n</i> <sup>3</sup> | $\sqrt[3]{n}$ | $\sqrt[3]{10n}$ | $\sqrt[3]{100n}$ |
|----------|-----------------------|------------|--------------|-----------------------|---------------|-----------------|------------------|
| 1        | 1                     | 1.000 000  | 3.162 278    | 1                     | 1.000 000     | 2.154 435       | 4.641 589        |
| 2        | 4                     | 1.414 214  | 4.472 136    | 8                     | 1.259 921     | 2.714 418       | 5.848 035        |
| 3        | 9                     | 1.732 051  | 5.477 226    | 27                    | 1.442 250     | 3.107 233       | 6.694 330        |
| 4        | 16                    | 2.000 000  | 6.324 555    | 64                    | 1.587 401     | 3.419 952       | 7.368 063        |
| 5        | 25                    | 2.236 068  | 7.071 068    | 125                   | 1.709 976     | 3.684 031       | 7.937 005        |
| 6        | 36                    | 2.449 490  | 7.745 967    | 216                   | 1.817 121     | 3.914 868       | 8.434 327        |
| 7        | 49                    | 2.645 751  | 8.366 600    | 343                   | 1.912 931     | 4.121 285       | 8.879 040        |
| 8        | 64                    | 2.828 427  | 8.944 272    | 512                   | 2.000 000     | 4.308 869       | 9.283 178        |
| 9        | 81                    | 3.000 000  | 9.486 833    | 729                   | 2.080 084     | 4.481 405       | 9.654 894        |
| 10       | 100                   | 3.162 278  | 10.00000     | 1 000                 | 2.154 435     | 4.641 589       | 10.00000         |
| 11       | 121                   | 3.316 625  | 10.48809     | 1 331                 | 2.223 980     | 4.791 420       | 10.32280         |
| 12       | 144                   | 3.464 102  | 10.95445     | 1 728                 | 2.289 428     | 4.932 424       | 10.62659         |
| 13       | 169                   | 3.605 551  | 11.40175     | 2 197                 | 2.351 335     | 5.065 797       | 10.91393         |
| 14       | 196                   | 3.741 657  | 11.83216     | 2 744                 | 2.410 142     | 5.192 494       | 11.18689         |
| 15       | 225                   | 3.872 983  | 12.24745     | 3 375                 | 2.466 212     | 5.313 293       | 11.44714         |
| 16       | 256                   | 4.000 000  | 12.64911     | 4 096                 | 2.519 842     | 5.428 835       | 11.69607         |
| 17       | 289                   | 4.123 106  | 13.03840     | 4 913                 | 2.571 282     | 5.539 658       | 11.93483         |
| 18       | 324                   | 4.242 641  | 13.41641     | 5 832                 | 2.620 741     | 5.646 216       | 12.16440         |
| 19       | 361                   | 4.358 899  | 13.78405     | 6 859                 | 2.668 402     | 5.748 897       | 12.38562         |
| 20       | 400                   | 4.472 136  | 14.14214     | 8 000                 | 2.714 418     | 5.848 035       | 12.59921         |
| 21       | 441                   | 4.582 576  | 14.49138     | 9 261                 | 2.758 924     | 5.943 922       | 12.80579         |
| 22       | 484                   | 4.690 416  | 14.83240     | 10 648                | 2.802 039     | 6.036 811       | 13.00591         |
| 23       | 529                   | 4.795 832  | 15.16575     | 12 167                | 2.843 867     | 6.126 926       | 13.20006         |
| 24       | 576                   | 4.898 979  | 15.49193     | 13 824                | 2.884 499     | 6.214 465       | 13.38866         |
| 25       | 625                   | 5.000 000  | 15.81139     | 15 625                | 2.924 018     | 6.299 605       | 13.57209         |
| 26       | 676                   | 5.099 020  | 16.12452     | 17 576                | 2.962 496     | 6.382 504       | 13.75069         |
| 27       | 729                   | 5.196 152  | 16.43168     | 19 683                | 3.000 000     | 6.463 304       | 13.92477         |
| 28       | 784                   | 5.291 503  | 16.73320     | 21 952                | 3.036 589     | 6.542 133       | 14.09460         |
| 29       | 841                   | 5.385 165  | 17.02939     | 24 389                | 3.072 317     | 6.619 106       | 14.26043         |
| 30       | 900                   | 5.477 226  | 17.32051     | 27 000                | 3.107 233     | 6.694 330       | 14.42250         |
| 31       | 961                   | 5.567 764  | 17.60682     | 29 791                | 3.141 381     | 6.767 899       | 14.58100         |
| 32       | 1 024                 | 5.656 854  | 17.88854     | 32 768                | 3.174 802     | 6.839 904       | 14.73613         |
| 33       | 1 089                 | 5.744 563  | 18.16590     | 35 937                | 3.207 534     | 6.910 423       | 14.88806         |
| 34       | 1 156                 | 5.830 952  | 18.43909     | 39 304                | 3.239 612     | 6.979 532       | 15.03695         |
| 35       | 1 225                 | 5.916 080  | 18.70829     | 42 875                | 3.271 066     | 7.047 299       | 15.18294         |
| 36       | 1 296                 | 6.000 000  | 18.97367     | 46 656                | 3.301 927     | 7.113 787       | 15.32619         |
| 37       | 1 369                 | 6.082 763  | 19.23538     | 50 653                | 3.332 222     | 7.179 054       | 15.46680         |
| 38       | 1 444                 | 6.164 414  | 19.49359     | 54 872                | 3.361 975     | 7.243 156       | 15.60491         |
| 39       | 1 521                 | 6.244 998  | 19.74842     | 59 319                | 3.391 211     | 7.306 144       | 15.74061         |
| 40       | 1 600                 | 6.324 555  | 20.00000     | 64 000                | 3.419 952     | 7.368 063       | 15.87401         |
| 41       | 1 681                 | 6.403 124  | 20.24846     | 68 921                | 3.448 217     | 7.428 959       | 16.00521         |
| 42       | 1 764                 | 6.480 741  | 20.49390     | 74 088                | 3.476 027     | 7.488 872       | 16.13429         |
| 43       | 1 849                 | 6.557 439  | 20.73644     | 79 507                | 3.503 398     | 7.547 842       | 16.26133         |
| 44       | 1 936                 | 6.633 250  | 20.97618     | 85 184                | 3.530 348     | 7.605 905       | 16.38643         |
| 45       | 2 025                 | 6.708 204  | 21.21320     | 91 125                | 3.556 893     | 7.663 094       | 16.50964         |
| 46       | 2 116                 | 6.782 330  | 21.44761     | 97 336                | 3.583 048     | 7.719 443       | 16.63103         |
| 47       | 2 209                 | 6.855 655  | 21.67948     | 103 823               | 3.608 826     | 7.774 980       | 16.75069         |
| 48       | 2 304                 | 6.928 203  | 21.90890     | 110 592               | 3.634 241     | 7.829 735       | 16.86865         |
| 49       | 2 401                 | 7.000 000  | 22.13594     | 117 649               | 3.659 306     | 7.883 735       | 16.98499         |
| 50       | 2 500                 | 7.071 068  | 22.36068     | 125 000               | 3.684 031     | 7.937 005       | 17.09976         |

# SQUARES, CUBES AND ROOTS

| $n$ | $n^2$  | $\sqrt{n}$ | $\sqrt{10n}$ | $n^3$     | $\sqrt[3]{n}$ | $\sqrt[3]{10n}$ | $\sqrt[3]{100n}$ |
|-----|--------|------------|--------------|-----------|---------------|-----------------|------------------|
| 50  | 2 500  | 7.071 068  | 22.36068     | 125 000   | 3.684 031     | 7.937 005       | 17.09976         |
| 51  | 2 601  | 7.141 428  | 22.58318     | 132 651   | 3.708 430     | 7.989 570       | 17.21301         |
| 52  | 2 704  | 7.211 103  | 22.80351     | 140 608   | 3.732 511     | 8.041 452       | 17.32478         |
| 53  | 2 809  | 7.280 110  | 23.02173     | 148 877   | 3.756 286     | 8.092 672       | 17.43513         |
| 54  | 2 916  | 7.348 469  | 23.23790     | 157 464   | 3.779 763     | 8.143 253       | 17.54411         |
| 55  | 3 025  | 7.416 198  | 23.45208     | 166 375   | 3.802 952     | 8.193 213       | 17.65174         |
| 56  | 3 136  | 7.483 315  | 23.66432     | 175 616   | 3.825 862     | 8.242 571       | 17.75808         |
| 57  | 3 249  | 7.549 834  | 23.87467     | 185 193   | 3.848 501     | 8.291 344       | 17.86316         |
| 58  | 3 364  | 7.615 773  | 24.08319     | 195 112   | 3.870 877     | 8.339 551       | 17.96702         |
| 59  | 3 481  | 7.681 146  | 24.28992     | 205 379   | 3.892 996     | 8.387 207       | 18.06969         |
| 60  | 3 600  | 7.745 967  | 24.49490     | 216 000   | 3.914 868     | 8.434 327       | 18.17121         |
| 61  | 3 721  | 7.810 250  | 24.69818     | 226 981   | 3.936 497     | 8.480 926       | 18.27160         |
| 62  | 3 844  | 7.874 008  | 24.89980     | 238 328   | 3.957 892     | 8.527 019       | 18.37091         |
| 63  | 3 969  | 7.937 254  | 25.09980     | 250 047   | 3.979 057     | 8.572 619       | 18.46915         |
| 64  | 4 096  | 8.000 000  | 25.29822     | 262 144   | 4.000 000     | 8.617 739       | 18.56636         |
| 65  | 4 225  | 8.062 258  | 25.49510     | 274 625   | 4.020 726     | 8.662 391       | 18.66256         |
| 66  | 4 356  | 8.124 038  | 25.69047     | 287 496   | 4.041 240     | 8.706 588       | 18.75777         |
| 67  | 4 489  | 8.185 353  | 25.88436     | 300 763   | 4.061 548     | 8.750 340       | 18.85204         |
| 68  | 4 624  | 8.246 211  | 26.07681     | 314 432   | 4.081 655     | 8.793 659       | 18.94536         |
| 69  | 4 761  | 8.306 624  | 26.26785     | 328 509   | 4.101 566     | 8.836 556       | 19.03778         |
| 70  | 4 900  | 8.366 600  | 26.45751     | 343 000   | 4.121 285     | 8.879 040       | 19.12931         |
| 71  | 5 041  | 8.426 150  | 26.64583     | 357 911   | 4.140 818     | 8.921 121       | 19.21997         |
| 72  | 5 184  | 8.485 281  | 26.83282     | 373 248   | 4.160 168     | 8.962 809       | 19.30979         |
| 73  | 5 329  | 8.544 004  | 27.01851     | 389 017   | 4.179 339     | 9.004 113       | 19.39877         |
| 74  | 5 476  | 8.602 325  | 27.20294     | 405 224   | 4.198 336     | 9.045 042       | 19.48695         |
| 75  | 5 625  | 8.660 254  | 27.38613     | 421 875   | 4.217 163     | 9.085 603       | 19.57434         |
| 76  | 5 776  | 8.717 798  | 27.56810     | 438 976   | 4.235 824     | 9.125 805       | 19.66095         |
| 77  | 5 929  | 8.774 964  | 27.74887     | 456 533   | 4.254 321     | 9.165 656       | 19.74681         |
| 78  | 6 084  | 8.831 761  | 27.92848     | 474 552   | 4.272 659     | 9.205 164       | 19.83192         |
| 79  | 6 241  | 8.888 194  | 28.10694     | 493 039   | 4.290 840     | 9.244 335       | 19.91632         |
| 80  | 6 400  | 8.944 272  | 28.28427     | 512 000   | 4.308 869     | 9.283 178       | 20.00000         |
| 81  | 6 561  | 9.000 000  | 28.46050     | 531 441   | 4.326 749     | 9.321 698       | 20.08299         |
| 82  | 6 724  | 9.055 385  | 28.63564     | 551 368   | 4.344 481     | 9.359 902       | 20.16530         |
| 83  | 6 889  | 9.110 434  | 28.80972     | 571 787   | 4.362 071     | 9.397 796       | 20.24694         |
| 84  | 7 056  | 9.165 151  | 28.98275     | 592 704   | 4.379 519     | 9.435 388       | 20.32793         |
| 85  | 7 225  | 9.219 544  | 29.15476     | 614 125   | 4.396 830     | 9.472 682       | 20.40828         |
| 86  | 7 396  | 9.273 618  | 29.32576     | 636 056   | 4.414 005     | 9.509 685       | 20.48800         |
| 87  | 7 569  | 9.327 379  | 29.49576     | 658 503   | 4.431 048     | 9.546 403       | 20.56710         |
| 88  | 7 744  | 9.380 832  | 29.66479     | 681 472   | 4.447 960     | 9.582 840       | 20.64560         |
| 89  | 7 921  | 9.433 981  | 29.83287     | 704 969   | 4.464 745     | 9.619 002       | 20.72351         |
| 90  | 8 100  | 9.486 833  | 30.00000     | 729 000   | 4.481 405     | 9.654 894       | 20.80084         |
| 91  | 8 281  | 9.539 392  | 30.16621     | 753 571   | 4.497 941     | 9.690 521       | 20.87759         |
| 92  | 8 464  | 9.591 663  | 30.33150     | 778 688   | 4.514 357     | 9.725 888       | 20.95379         |
| 93  | 8 649  | 9.643 651  | 30.49590     | 804 357   | 4.530 655     | 9.761 000       | 21.02944         |
| 94  | 8 836  | 9.695 360  | 30.65942     | 830 584   | 4.546 836     | 9.795 861       | 21.10454         |
| 95  | 9 025  | 9.746 794  | 30.82207     | 857 375   | 4.562 903     | 9.830 476       | 21.17912         |
| 96  | 9 216  | 9.797 959  | 30.98387     | 884 736   | 4.578 857     | 9.864 848       | 21.25317         |
| 97  | 9 409  | 9.848 858  | 31.14482     | 912 673   | 4.594 701     | 9.898 983       | 21.32671         |
| 98  | 9 604  | 9.899 495  | 31.30495     | 941 192   | 4.610 436     | 9.932 884       | 21.39975         |
| 99  | 9 801  | 9.949 874  | 31.46427     | 970 299   | 4.626 065     | 9.966 555       | 21.47229         |
| 100 | 10 000 | 10.00000   | 31.62278     | 1 000 000 | 4.641 589     | 10.00000        | 21.54435         |

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